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Train Communications

During the past few years many railroad companies have conducted extensive tests of inductive and radio carrier communication systems. Excerpts from reports of these tests point out that these systems though not perfect are already rendering useful service.

ALTHOUGH radio communication for use by railroads was conceived several years ago,[1] it has not come into extensive use for several reasons, among which is the hesitancy to adopt any system until all adverse operating characteristics are eliminated. The Association of American Railroads, the radio equipment manufacturers, and the Federal Communications Commission, with the co-operation of the major railroads of the United States, have conducted during the last few years very extensive tests of both inductive and radio carrier communications. These tests have been conducted under actual operating conditions by installations on regular established scheduled trains, and have formed the best possible basis for selection of a suitable communication system.

The following is a symposium from the reports on these tests.

The Atchison, Topeka and Sante Fe Railroad in June 1944 conducted tests of end-to-end communication on a 3,500-ton freight train on a run from Bakersfield, Calif., to Chicago, Ill. For these tests a modified turnstile type of antenna was mounted on the top of Diesel locomotives and in the case of steam locomotives, on the top of the tender. An antenna of the same type was mounted on the roof of a business car attached to the rear of the train. The reception at both ends was unaffected in traveling over all types of terrains, in mountains, deep cuts, grooves, and under steel over-structures. Varying atmospheric conditions, including two rain and electric storms, were encountered during the 2,200-mile trip, but had no effect on the communication. The maximum distance of transmission of about two miles was accomplished at times when the engine was uncoupled from the train and moved to a roundhouse or to service tracks. Quality and volume of transmission was satisfactory at all times except for background noise from the equipment located on the steam engine which, while not desirable, was not sufficiently objectionable to interfere with transmission.

The Denver and Rio Grande Western Railroad has conducted extensive tests of end-to-end and point-to-train communication in which radio was used on frequencies ranging from 30 to 2,700 megacycles. The following is part of the report to the Federal Communications Commission, dated December 8, 1944.

Despite the adverse conditions of terrain and extensive tunnels, the results of the tests to date are exceptionally encouraging. Early in the tests it became evident that the greatest source of

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Material for this article was obtained from the Association of American Railroads, New York, N. Y.

Figure 1. Inductive carrier station aboard a Seaboard Railway engine

communication difficulties would be in the tunnel area, a 41 mile stretch, where the Rio Grande operates over the Denver & Salt Lake Railway line, starting 15 miles west of Denver and extending through the Moffat Tunnel. In this territory there are 30 tunnels varying from a few hundred feet in length to the Moffat Tunnel which is 6.2 miles long.

In this territory due to the frequency of the tunnels there are numerous times that either the locomotive or caboose, or both, are in a tunnel at the same time. It is in this extremely rugged territory with difficult operating conditions that the use of radio as a reliable means of communication would introduce an added factor of safety and materially aid in the handling of trains and expediting the movement of traffic.

In conducting caboose-to-locomotive communication tests, there were attempts to use different types of equipment varying in receiver sensitivity, transmitter power, frequency, and method of modulation. All equipment is installed in freight service—one unit on a Diesel locomotive, the other on a caboose—and tests are made between front and rear of the freight train on the main line between Denver and Salt Lake City or some intermediate point.

The Northern Pacific Railway conducted tests on radio equipment during August 1945.

Although it was fully anticipated that some blind spots would be encountered due to curvature and foliage especially in mountain areas, the only dead spots we actually encountered were tunnels. Good communication was obtained everywhere except when one or both ends of the train were in a tunnel and in every case when the train was long enough or the tunnel short enough, communication was established around the tunnel as soon as the engine emerged and the caboose had not yet entered.

On October 17, 1945, end-to-end and point-to-train communication tests using both inductive carrier and radio equipment were conducted on the Illinois Central Railroad between Freeport, Ill., and Waterloo, Iowa.

Radio communication between the ends of the train could be effected at all times. As the train approached the radio station at Jesup, Iowa, attempts were made to communicate with this station at a distance of approximately 40 miles from the station without success. At points between 19 and 11 miles from Jesup the station could be heard intermittently, but consistent communication was not possible. Good communication was established at a point about 11 miles from Jesup and maintained from that point to about the same distance beyond Jesup.

As the train was approaching the Jesup station, arrangements were made for a special test as follows:

The station at Jesup was directed to transmit continuously for a considerable period. The radio station on the locomotive was directed to transmit also. Observations were made on the radio receiver on the test car at the rear end of the train to determine the effect when the signals from the Jesup station and those from the locomotive reached approximately the same intensity at the test receiver. Results were as follows:

As the rear end of the train passed the Jesup station, and for a distance of a mile or two beyond that point, the Jesup station maintained control of the observation receiver, and no signal or interference was heard from the radio transmitter on the locomotive which was about 3/4 mile ahead of the test car. At a point somewhere between one and two miles from the Jesup station the transmitter at the front end of the train took over control of the receiver. This was accomplished without any appreciable flutter or presence of both signals. The train, during this test, was moving at a fairly rapid speed. Although this test was very limited in scope, it seems probable that, in the case of a train moving at normal speeds, the area in which communication would be obscured by flutter because of approximately equal signal intensities received from two wayside stations will probably not be great.

During April 1945, the Texas and New Orleans Railroad Company conducted tests on the inductive type of communication, on end-to-end, point-to-train, and train-to-train very-high-frequency radio communication.

The first train started from El Paso, and made a run with tonnage freight to Valentine (161.5 miles east). On this trial run it was expected that satisfactory communication would be obtained at all times between the locomotive and caboose and with a fixed station whenever the train was not more than ten miles distant. The distance that the train did communicate with the station varied and under favorable conditions contact was made up to twenty miles between the train and one of the fixed stations. For this test the first train left Valentine for a return run to El Paso and the second train left El Paso for Valentine, it being the intention to test the radio equipment between two trains approaching, meeting, and leaving each other. The first satisfactory contact between the two trains was established when the trains were about six miles apart and was maintained continuously until the trains met and were again about five or six miles apart. The terrain between El Paso and Valentine is generally mountainous, the railroad usually traversing the low portion between high hills and mountains which affords a very severe test of radio equipment of the frequency used because of the general line of sight characteristics at this frequency.

The Seaboard Air Lines Railway conducted a test run in 1944 between Richmond, Va., and Savannah, Ga., using inductive equipment.

Transmission was loud, clear, and of good quality. When passing a 110-kv power line no interference was noticed.

No difficulty was encountered in going around curves, through cuts, and so forth. At one time the train passed through a steel truss bridge with steel members on the bottom side and top. This did not seem to affect transmission in the least.

An inductive radio test was made on the Southern Pacific in 1944 from moving train to fixed stations and from head to rear end, in which a 73-car freight train was operated over a distance of 180 miles. This train originated at Roseville, Calif., and ran through to Fresno. When running in parallel to the telegraph line carrying the Western Division wires we were able to communicate from moving train to Stockton (42 miles). The volume and quality improved as the train proceeded to Stockton; however, it was audible and understandable at all times. The country through which we traveled was flat and level and telegraph line varied in distance from center line of the track from 50 to 300 feet, but the greater portion of the line averaged about 50 feet from center line of the track. The maximum distance that we talked from moving train to fixed station was some 70 odd miles. The test was quite satisfactory in every respect. We experienced some slight interference where wig-wags were operating in yards, but these noise levels did not interfere with the voice to the extent that it could not be clearly understood.

The Chicago, Milwaukee, Saint Paul, and Pacific Railroad, during December 1944, conducted tests of end-to-end and point-to-train communication using the inductive system.

Diesel locomotive and a mail and express car were equipped with mobile sets. At the Beloit Division office a similar set was installed.

On the westbound trip communication was carried on with Beloit satisfactorily from about 50 to 60 miles east to about 20 to 30 miles west of that point. Eastbound the range was from 30 to 40 miles west to 60 miles east. Electrical interference attributed to the auxiliary and main generators of the locomotive interfered seriously with the reception on the locomotive at certain engine speeds and loads. The effect of this interference was to render two-way communication practically impossible. The application of filters to certain parts of the locomotive electrical equipment is planned.

The New York Central Railroad during April 1945, conducted tests of end-to-end and point-to-train inductive carrier communication.

These installations operated on 175 kc, employing frequency modulation. Loudspeakers were provided for calling and handsets for talking and listening although both the handset receiver and loudspeaker could be used simultaneously by depressing the handset hook.

Transmission between all fixed stations was of good quality, quiet, and of ample volume. Communication between the mobile units was also good except that when wayside wires were below track level communication, particularly to and from the engine, was noisy. On one test when the engine was uncoupled from the train communication between engine and caboose was carried on over a distance of six miles. Communication between the mobile units and fixed stations was also satisfactory; the maximum distance over which such communication was attempted was 27 miles.

INDUCTIVE COMMUNICATION

Inductive train communication consists of communicating between mobile units (locomotives and cabooses) and between mobile units and fixed stations by means of carrier currents which are superimposed on the usual telegraph and telephone wires which parallel the right of way. Communication is by telephone, two-way, usually simplex operation, and as far as the user is concerned is similar to radio. One or more than one channel can be provided. It in no way interferes with the normal use of these telegraph and telephone wires.

For mobile units coupling is usually obtained by means of vertical plane loop antennas in line with the track. On transmitting, these loops are energized with the carrier frequency modulated by the voice signals, and induce, by electromagnetic induction, such currents longitudinally into all wires on the pole line. Thus these induced currents flow in all wires and appear as a voltage between all wires and ground. On receiving the carrier currents flowing in the line are induced into the loop. Some systems use the same loop for both transmitting and receiving, others use a separate smaller loop for receiving, and if there are two channels, use a separate receiving loop for each frequency.

At the fixed station the set is connected between ground and one or more wires on the pole line through a coupling unit, which consists essentially of small capacity condensers. While coupling is usually made to one wire, or to the two wires of a telephone circuit only to keep the pair balanced for telephone service, the capacity between wires on the pole line for these high frequencies, makes this essentially coupling to all wires on the pole line. Tests have indicated there is no advantage gained coupling the way station to more than one or two wires.

On some tests the coupling on mobile units has been made by means of a capacity-type antenna, and coupling to the line wires obtained by electric induction. However, for transmitting the loop antenna has become almost universal, while the capacitance-type antenna is used in some cases for receiving.

Certain of the earlier systems used the upper side band of a 5,700-cycle carrier. Amplitude modulation at 75 and 80 kc has been used and frequency modulation from 70 to 200 kc. The usual frequency modulation deviation ratio used is one-to-one, which occupies a band about 6 kc wide 3 kc above and below the carrier frequency. Manufacturers state that they can supply equipment to operate from 50 kc to 250 kc, and apply frequency modulation.

Frequency modulation, and the use of the higher carrier frequencies, has the advantage of reducing extraneous noise.

By far the greatest loss is the coupling loss between the mobile units and the pole line. It will increase with the separation of the pole line from the track, and if the pole line is above or below the level of the antenna on the mobile units. The coupling loss is less the higher the frequency. Data are very meager and because of the large number of variables would be difficult to set up. Data obtained from one manufacturer, at 170 kc, show the coupling loss between one mobile unit and the pole line to be between 25 and 45 decibels for separations from 30 to 200 feet.

The coupling loss between the fixed station and the line is practically zero.

The loss along the pole line will increase with the frequency; representative values given were 0.4 decibel per mile at 70 kc and 0.8 decibel per mile at 170 kc, although on some lines it has measured as high as 1.2 decibels per mile. It will be increased to a certain extent by the presence of iron wire and bridged taps, particularly if these bridged taps are long and go into lead-sheathed or underground cable. The effect of bridged taps can be eliminated by the use of blocking filters and in some cases this may be necessary.

The Kansas City Southern Railway demonstrated the use of carrier telephone equipment on June 21, 1944.

The system is a carrier system and 80 and 175 kc were used in operation. The system extends from Kansas City, Mo., to Shreveport, La. (560 miles). At present one engine and two cabooses are equipped. Shortly it is expected five engines and five cabooses will be equipped. There are 17 fixed stations along the line from two to 62 miles apart. The system is used for station to station, station to train, and front to rear end communication. The carrier is superimposed upon a telephone pair which, in addition, provides a voice channel, one simplex telegraph channel, and a 22-kc telephone and telegraph carrier system, the latter providing one long distance telephone and two 2-way telegraph channels.

The set on the engine is connected to a loop consisting of seven turns of number 12 copper wire. During the test the loop used was mounted on top of the tender in a vertical plane and was approximately six feet high and 12 feet long. At other times a loop arrangement had been tried on the engine consisting of one turn on each side of the boiler. The top wire was supported about one foot away from the boiler and about on a level with the sand dome. The bottom wire is just below the runway. These two loops, one on each side of the engine, were connected in series to give the effect of a 2-turn loop.

So far as reception and transmission are concerned, either arrangement appeared to work satisfactorily.

The loop on the caboose consisted of four turns. All four turns were carried across the top of the caboose in the center. On the bottom two of the turns were carried on each side and just below the framework of the caboose.

The transmitter output of 75 watts at carrier frequency is fed into the loop of either the engine or the caboose.

The system operates from the fixed station by inducing a magnetic field around the paralleling wire line which cuts the turns of the loops on the moving equipment and induces voltages therein. These voltages feed into the receivers on the moving equipment.

On the moving equipment the transmitters circulate currents through the loop and these currents generate magnetic fields which in turn cut the wire line and induce voltages therein. At the fixed station these voltages act directly on the receiver. They also produce current in the paralleling line wire and this current generates a second field which interacts with the loops on other units of mobile equipment.

The arrangement between fixed station and mobile equipment is somewhat analogous to a 2-winding transformer. The arrangement between mobile units is somewhat analogous to a 3-winding transformer, winding number 1 corresponding, say, to a caboose, winding number 2 to the paralleling wire line and winding number 3 to the equipment on the engines.

One company has advised that they have found that when the loops are oriented in the vertical plane they provide the most efficient transfer of energy to the wire line when the latter is centered and perpendicular to the loop. For a satisfactory transmission the wire line should be not more than 125 feet away from the track.

RADIO COMMUNICATIONS

The technical methods and equipment used varied from company to company. The Denver and Rio Grande Western Railroad reports

To date four different types of radio equipment have been tested in actual service tests:

1. Frequency-modulation 60-watt emergency equipment operated in the 35-megacycle band.
2. Phase-modulation 15-watt experimental unit of the Navy in the 2,600-megacycle band.
3. Frequency-modulation 25-watt repeater equipment on a frequency of 117.65 megacycles.
4. Air-bourne amplitude modulation 6 watt *SCR-624A* unit on a frequency of 156.525 megacycles.

A number of antenna arrangements have been attempted. The vertical quarter wave to ground, vertical folded dipole, horizontally polarized dipole, *J* type with matching stub and compensating capacitance, and an array composed of a driven vertical quarter wave with a parasitic reflector and director.

The 60-watt frequency modulation emergency equipment proved very satisfactory everywhere except in the Moffat Tunnel. The test was conducted on a freight train, caboose-to-locomotive two-way from Denver to Salt Lake City and return. We did not encounter any noise or interference from man-made static. No fading, flutter or distortion was noticed throughout the trip. The 15-watt phase-modulation experimental unit was a one-way, locomotive-to-caboose test, mounted on a freight train, and tested from Denver to Salt Lake City. It was possible to hold conversation in one direction, but as the system of testing had to be interrupted, it was not deemed advisable. Field strength was measured at the receiver and a continuous wave audio note was modulated on the carrier for the benefit of a loud speaker in the caboose. Because of a radio equipment failure, and the brief time the equipment was available to us, no test was obtained in the "shelf" area where the track runs for some distance on the side of a mountain with no mountains or other possible reflecting surfaces on the other side of the track. It is possible that this area could give trouble at this frequency. No flutter or distortion was apparent. Fading, of course, was registered on the meter, and in a few cases went below the level where the automatic voltage control could maintain normal speaker volume. In one case, with the locomotive in a tunnel, the signal died because of lack of reflecting surface. It is thought that a parasitic reflector could be used to re-radiate the wave. An interesting fact was that radiation was more effective in the six mile long Moffat Tunnel than out of it, the tunnel apparently acting as a wave guide. No flutter or distortion was apparent throughout the test.

The 25-watt frequency-modulation repeater equipment did not perform in any of the tunnels, and in some cases we had trouble obtaining enough volume in cuts, canyons, and so forth, to properly operate the speakers. This trouble was no doubt due to the fact that this equipment uses such a low-grain receiver. No flutter or distortion was evident except when we used a horizontally polarized dipole antenna. No noise or interference from man-made static was encountered. All tests with this equipment were conducted on a freight train, engine-to-caboose, two-way between Denver and Grand Junction, Colorado.

The air-bourne units are now being tested in freight service, engine-to-caboose two-way between Denver and Grand Junction. No reception has been obtained in the tunnels, consequently giving us very poor service in the tunnel area. The only time interference is noticed is when the units are in close proximity to motor cars, probably due to the magnetos used for ignition. Under this condition some flutter becomes evident but the signal is still intelligible.

In comparing the signals received by the different units, the re-

ceivers used in each case must be considered. The sensitivity of all the receivers used is of the order of three microvolts for a ten decibel signal-to-noise ratio, except in the case of the 25-watt repeater units which have a rather low gain.

In the case of the antenna experiments, the best results to date have been obtained from the use of the J type and the quarter-wave vertical to ground. We do not consider the horizontal type of antenna satisfactory for this type of radio communication. The other types of antennas used varied somewhat in the signal strength received, but no harmful effects were noticed.

Table I. Railroad Radio Authorization by Federal Communications Commission (to July 15, 1946)

Railroad	Transmitter Location	Nature of Installation
The Atchison, Topeka and Santa Fe	Chicago, Ill.	Yard—fixed station
	San Francisco, Calif.	Yard—fixed station
	(Portable and portable-mobile)	14 stations
	Chicago, Ill.	Fixed station
The Baltimore and Ohio	Newcastle, Pa.	Yard—3 units
Chicago, Burlington and Quincy	Chicago, Ill.	Yard—fixed station
	(Mobile)	Train stations—30 units
Chicago, Milwaukee, Saint Paul and Pacific	(Mobile)	Train stations—15 units
Chicago, Rock Island and Pacific	(Portable and portable-mobile)	Experimental
	(Portable)	Train stations—7 units
	(Portable)	Yard and Terminal
	(Mobile)	14 units
	(Mobile)	Train stations—20 units
The Denver and Rio Grande Western	(Portable and portable-mobile)	Experimental
	(Mobile)	Train stations—32 units
	Denver, Colo.	Yard and terminal—fixed station
	(Mobile)	Yard and terminal—7 units
Florida East Coast Railway Company	Buena Vista Railway Yards, Miami, Fla.	Yard and terminal—fixed station
	(Mobile)	Yard and terminal—11 units
Gulf, Mobile & Ohio	Meridian, Miss.	Yard & terminal—fixed staton
Jacksonville Terminal Company	Lee St. Tower, Jacksonville, Fla.	Terminal—fixed station
	Myrtle Ave. Tower, Jacksonville, Fla.	Terminal—fixed station
	(Portable-mobile in vicinity of Jacksonville)	Terminal stations—12 units
Missouri Pacific Railroad	(Mobile)	Train stations—32 units
New York Central	Cheektowaga Township, N. Y.	Yard—fixed station
	Bethlehem Township, N. Y.	Yard—fixed station
	Manlius Township, N. Y.	Yard—fixed station
	(Portable and portable-mobile)	Yard—24 stations
	(Portable and portable-mobile)	Yard—12 stations
	Hammond, Ind.	Yard—fixed station
	(Mobile)	Yard—5 units
Northern Pacific Railway Company	(Mobile)	Train stations—5 units
Pere Marquette Railway Company	Grand Rapids, Mich.	Train Station—fixed
	(Mobile)	Train stations—5 units
Seaboard Air Line Railroad Company	Hamlet, N. C.	Yard and terminal—fixed station
	Hamlet, N. C.	Yard and terminal—fixed station
	(Mobile)	Yard and terminal—fixed station, 15 mobile units
Union Pacific Railroad Company	Fairfax Railroad Yard, Kansas City, Kans.	Yard and terminal—fixed station
	(Mobile)	Yard and terminal—16 units

Train radio tests were conducted on the New York, New Haven, and Hartford Railroad in the period September 12 to December 15, 1944, utilizing frequency modulated equipment operating at a frequency of 30.66 megacycles. Three sets were constructed. In its final form, each set of equipment consisted of a transmitter, a power supply, and a receiver, shock mounted in a dust-tight weatherproof steel housing. A handset microphone of the push-to-talk type was used in each transmitting station. Voice reception was by either the handset or a loud speaker, the loud speaker serving to announce the initiation of a call.

The two mobile stations had an output of 30 watts each and the fixed station 60 watts.

The antennas on the mobile equipment were in the form of a U-shaped pipe railing mounted on the tender just behind the coal pile and on the caboose roof alongside the cupola. In all cases the antenna was entirely within the clearance limitations of the equipment and was effectively grounded to provide protection from possible accidental contact with overhead trolley wires in the electrified territory. For the fixed station, an antenna in the form of a large ring located on top of a grounded steel pole was used.

It was found that the range of the radio equipment was not fixed. It varied with the height of the antenna, condition of terrain, and other factors which caused shielding of the radio signal. However, based on tests at several locations, it was found that satisfactory operation could be obtained between the fixed station and mobile units up to distances of approximately 10 miles and between mobile units at separations up to approximately 5 miles.

RELAY SYSTEMS

It is likely that radio relay systems will find application in the railroad communication field where there is a sufficient volume of traffic to justify the cost of installation or where unusual conditions cause frequent interruptions to continuity of service rendered by wire lines and the cost of maintenance of such lines is unusually high. It should be pointed out, however, that it now appears the cost of radio relay systems will be so high that they probably cannot be economically justified except in situations where they can be operated with fairly heavy loads.

In the railroad field it appears that there may be a possibility of operating relay systems in conjunction with train-to-point radio systems where it will be possible to utilize common equipment for power supply, antenna supporting structures, auxiliary power units and housing facilities, which will tend to make for more economic installations of both types of equipment and may make possible the utilization of radio relay links under conditions where traffic loads would not ordinarily justify such installations.

So far as is known, the only railroad installation to date of point-to-point communication by use of ultra-high frequency is an experimental link set up between Kansas City, Mo., and Topeka, Kans., by the Rock Island Lines.

EQUIPMENT

As a result of the subject tests equipment designed for railroad use in the 158–162-megacycle band has been

developed by several manufacturers. Equipment used in the earlier tests was police, aviation, and military radio equipment adapted for test use in connection with railroad operations. As might be expected some failures of this equipment occurred. Attempts have been made in the new equipment to eliminate the weaknesses exposed and develop rugged equipment capable of dependable operation in railroad service. One requirement made clearly evident by the tests was the need for sensitive receivers in order that satisfactory communication may be obtained where mountains, deep cuts, or foliage, may cause severe attentuation of received signals.

Table II. Induction Installations

Date in Service	Railroad	Type Equipment
Mar. 1937	Bessemer and Lake Erie	2-way
Mar. 1940		
July 1940	Cleveland, Cincinnati Chicago and Saint Louis	1-way
May 1943	Chicago, Burlington and Quincy	1-way, 2-freq.
Mar. 1944	Chicago, Burlington and Quincy	1-way
May 1943	Great Northern	1-way
Oct. 1940	Louisville and Nashville	1-way
Apr. 1942	Norfolk and Western	2-way
Mar. 1943	Pennsylvania Railroad	1-way, 2-freq.
Dec. 1942	Pennsylvania Railroad	2-way
Nov. 1940	Pennsylvania Railroad	1-way
May 1942		
June 1943	Pennsylvania Railroad	1-way, 2-freq.
Mar. 1942	Pennsylvania Railroad	1-way
July 1943	Pennsylvania Railroad	1-way
Aug. 1944	Pennsylvania Railroad	2-way
Feb. 1945	Terminal Railroad Association of Saint Louis	1-way
	Kansas City Southern Railroad	2-way
(Proposed)	Atlantic Coast Line	2-way, 2-freq.
(Proposed)	Chesapeke and Ohio	2-way
(Proposed)	Pennsylvania Railroad	2-way, 2-freq.

Adequate automatic control of receiver output volume is another important requirement.

The latest transmitting equipment developed for this service generally has a power output of from 5 to 50 watts, radio frequency range of 152 to 162 megacycles, audio frequency range of 300 to 3,000 cycles, and is usually designed for use with 52- or 72-ohm loads.

The inductive carrier equipment, having been under development for railroad use over a longer period, has been standardized to a greater degree. Power output of this equipment is uniformly 5 watts for wayside stations and 50 watts for mobile stations, employs carrier frequencies between 88 and 175 kc, and frequency modulation with a 1-to-1 deviation ratio.

Radio equipment used in the various tests employed both amplitude and frequency modulation. It was found during these tests that noise from electrical sources caused very little trouble in reception of signals above 150 megacycles. While it was apparent that frequency modulation resulted in greater exclusion of interference from other stations on the same frequency, and with amplitude modulation heterodyne interference occurred, there has been insufficient operating experience in regular service to determine whether this discrimination characteristic of frequency modulation would be an advantage or disadvantage.

Little difficulty has been experienced in teaching railroad operating personnel to use either radio or inductive carrier equipments. In general the equipment has a minimum of external controls. Operation of a "push to talk" button on the microphone is the only control used in normal operation of the equipment.

The main problem in connection with antennas for both radio and inductive carrier has been to develop efficient antennas for mobile use within the limitation as to height imposed by the railroad clearance pattern. On frequencies above 150 megacycles horizontal antennas caused variation or "flutter" in the received signals. Consequently most radio antennas for mobile service have been vertical quarter-wave or variations thereof. The most used variations are top loaded vertical antennas, usually with a ground plane of the cartwheel type, and modifications of the coaxial type of antenna.

Fixed station antennas, although uniformly vertical, have varied rather widely in other respects. Some directional arrays have been used in attempts to extend the range of point-to-train radio communication. Considerable development work must be done in connection with fixed station radio antennas.

In inductive carrier communication, loop antennas, oriented in a vertical plane parallel to the tracks, have been used almost entirely. The particular form of loop varies, sometimes being entirely insulated from the mobile unit on which it is installed and sometimes being connected to the unit in such a manner that the output is fed inductively into the rails.

In end-to-end communication adequate volume and quality of signals have been maintained in the various tests with the exception of a few isolated instances where some flutter occurred in territory including deep cuts and foliage. The tests have indicated that radio communication in the 158–162 megacycle range fails in tunnels although some efforts have been made toward development of methods of overcoming this difficulty. One test involving tone-modulated 2,660-megacycle equipment was successfully received through the 6-mile Moffat tunnel but failed in a shorter tunnel because of lack of reflecting surfaces.

Inductive carrier communication in end-to-end service has been equally successful except where wayside wires are too far from the track or too far above or below track level. The maximum permissible distance between wayside wires and track averaged 200 feet. Greater attenuation was noted when wires were below track level than when considerably above track level.

Good communication over distances of approximately 5 miles by radio and approximately 15 miles by inductive carrier was obtained in end-to-end and train-to-train service. In point-to-train service the average distance

was 12 miles for radio and 50 miles for inductive carrier. Results of the tests show that line-of-sight conditions are generally necessary for reliable radio communication in the 158–162 megacycle band, although in some cases communication ranges well beyond line-of-sight are realized.

USES

End-to-end communication saves substantial amounts of time on many operations. Radio communication is not reserved for emergencies, but is used extensively in normal operations. For example

1. The engineer may be notified when proper train line pressure has been reached at the rear of the train.

2. The engineer may be notified as soon as the rear of the train clears crossovers and switches, thus permitting full speed sooner than would otherwise be possible.

3. Increased efficiency may be obtained in detecting and remedying hot boxes.

4. The engineer may be notified of the flagman's return to the train, thus saving time and eliminating uncertainty.

5. Control of Diesel-electric locomotives may be facilitated through more effective co-ordination in manipulating controls.

6. Unusual conditions may be discussed by the conductor and engineer.

7. The engineer may be advised of the time to start and stop at specific points.

8. The engineer may be advised of sticking brakes.

9. The conductor may advise the engineer to stop the train, thus avoiding hazards incident to application of air from rear of train. (Application of the airbrakes from the rear of the train, which might be necessary were it not for radio communication, may part the train and cause delay to other trains.)

10. The engineer may advise the conductor where water is to be taken, when a train to be met is in clear on the siding, or when difficulties develop with the engine.

11. The conductor may advise the engineer when the rear of the train is nearing switch so that the train may be stopped close to switch to pick up the trainman after lining the switch.

12. The conductor may advise the engineer when the brakes are released on the rear of the train.

13. The conductor may advise the engineer when to stop the train to spot cars to be loaded, unloaded, or re-iced.

14. The conductor is often able to notify the engineer when fire is dropping from firepan.

15. The engineer may notify the conductor of the indication of train order and interlocking signals.

16. If the train should break in two while in a siding the conductor may instruct the engineer as to the proper handling to couple.

17. If the conductor should miss a message handed up at a station, the engineer may read him the contents from his copy, or vice-versa.

Numerous advantages are also available from point-to-train communications.

1. The dispatcher may advise the train on what track cars are to be set out before the train arrives.

2. The conductor may advise the dispatcher when a train he is meeting leaves a station.

3. The dispatcher may inform the conductor of unscheduled train and designate the nearest siding.

4. The dispatcher may tell the crew what work is to be done when they arrive at station.

5. The conductor may advise the dispatcher of low steam and delay in schedule. Necessary orders may then be issued and repeated.

6. The dispatcher may advise the train where to wait to avoid being delayed by an opposing train.

7. The conductor may advise the dispatcher that the train has broken in two and is delayed.

8. The conductor may advise the dispatcher when the train has departed.

The Denver Rio Grande Western Railroad reports that radio communications aided them on their test run.

In one case, the radio train approached Grand Junction from the east and learned that the yards were congested. There are two ways of entering Grand Junction—one on the main line, and one from the south with switches from the main line about a mile and a half east of the main line entrance. The latter route leads to the east (or storage) yard, which is used for train handling only when the main yard is congested. The usual method of handling this situation is for the head end brakeman to walk back to the caboose and inform the conductor of the situation, and then return to the middle of the train to relay the hand signals from the rear to the front. The train is then backed up $1^1/_2$ miles into centralized traffic control territory to clear the switch which will take the train into the yards on the alternate route. This takes considerable time because of the walking involved and the caution with which the train must be moved back. In the case of the radio equipped train, the engineer informed the conductor immediately of the condition in the yards, the conductor ordered the train to back up, and he called signals over the radio. It is estimated that the radio saved about an hour and 15 minutes in handling the situation, besides avoiding the possible detention of any other train that might have been close behind.

On another westbound trip the centralized traffic control headed the train into the passing track at DeBeque, Colo. There was a broken rail between DeBeque and Akin and the dispatcher allowed trains to proceed on a permissive card only. We filled the passing track at DeBeque. There were four passenger trains behind us, and they could not proceed to Akin without "sawing" by the congested passing track. The dispatcher asked us to break our train and put the front half on a stock track adjoining the passing track so he could send the extra train at Akin to occupy our track with us. This would allow a clear track at Akin for the four passenger trains to go through. This was a complicated move and as our orders came from the telephone near the head end, the information had to be relayed back to the conductor on the caboose. His acknowledgment was received and instructions as to how to handle the cut given by him. The move was made without the conductor having to leave the caboose. No walking was involved except for the head brakeman to walk back 20 cars to effect the cut. The trainmen suggested that in this case probably an hour's time was saved by the use of the radio.

REFERENCE

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Science, Politics, and the National Welfare

FREDERICK L. HOVDE

MY AUDIENCE TODAY is composed of scientists, engineers, educators, and executives, all of whom are concerned professionally with that special and highly useful branch of science called "electronics." You are met to discuss the fate of this science, to speculate on its future, to discuss its problems, to exchange information, and to make plans enabling your science to increase its service to society.

Such conference objectives are laudable and worthy of your time and attention as active scientists. But you are also well-educated and responsible members of a society whose welfare depends primarily on how wisely it is governed and how well it uses the fruits of your work. Here is a problem far more important and far more difficult than any technical problems you will discuss in the conference sessions which follow.

Originally presented as an address before the recent National Electronics Conference in Chicago, this article considers one viewpoint in the current controversy over the question of government support of scientific research. The author takes his stand on the side of federal promotion of science, provided the policy to be pursued is both scientifically and politically acceptable. Undoubtedly, some readers will take issue with him. Truly, here is food for serious thought, and potential for extended debate.

This problem must, I believe, receive your constant and serious attention; otherwise your scientific labors will be wasted—even worse, will be used to destroy a civilization potentially finer than any in the past because man at long last has the power to control the forces of nature. Highly doubtful, but more important, is the question of whether or not man has the power and will to control himself.

UNION OF SCIENCE AND POLITICS

Sigmund Freud, the great Viennese scientist, founder of psychoanalysis and contributor to modern psychiatry and psychology, once remarked that man had three impossible tasks—namely, to educate, to govern, and to heal. This pessimistic observation is uncomfortably near the truth, as a little reflection on the history of man's progress with these tasks will show.

We cannot, however, accept Freud's dictum. To accept defeat before the battle is to be defeated for sure. We must continue relentlessly with all the genius and hard work at our command to attempt the solution of the problems of educating, governing, and healing. Even if Freud was right, we must continue our search for the answers. The intellectual and spiritual activity involved is, by its very nature, justification enough.

In the title of this article I have coupled the words "science" and "politics." These bedfellows are strangers. Some of you may think that they are incompatible, and, therefore, not likely to produce much. Whatever your thoughts about this union, they are joined together for better or for worse in both our national and our international life. The scientist is perhaps more surprised than the politician, who long has been used to strange bedfellows. In preparing these remarks I thought not only of the commonly accepted meanings of the words of my title, but also of their dictionary definitions. In Webster's dictionary you will find that "science" comes from the Latin word meaning "to know" and its definitions is "accumulated and accepted knowledge, as of principles or facts, systemized and formulated with reference to the discovery of general truths or the operation of general laws." The word "politics" comes from the Greek word meaning "citizen." Webster gives its meaning as "the art of government; the theory or practice of managing affairs of public policy or political parties."

If we accept these definitions (and Webster is hard to improve upon) then our bedfellows are not so incompatible. Indeed, one wonders why it has taken them so long to court each other. Actually they are not incompatible—they have lived together in separate worlds and just had not been introduced to each other. Never having been introduced, they could not be expected to understand each other.

It took World War II to introduce them. For the first time in the history of the world, there has been a full partnership between the government and the forces of science and technology. This partnership revolutionized military strategy and problems of national defense.

Full text of an address presented at the National Electronics Conference, Chicago, Ill., October 3–5, 1946.

Frederick L. Hovde is president, Purdue University, Lafayette, Ind.

The fruits of this partnership now are known to everyone. They include radar, rockets, jet-propelled aircraft, guided missiles, proximity fuses, and, of course, the acme of all destructive weapons, the atomic bomb. In all of these, your science of electronics is concerned. Further knowledge in the fields of chemical warfare and bacteriological warfare is certainly hidden in our laboratories. The war through which we have just passed was indeed a total war, for even the psychologist and the economist played important parts and put their knowledge to use in the conflict. We saw the results of carefully planned and well-organized operations in both psychological and economic warfare. Fortunately, almost all the scientific developments of war are useful for the peace. The same knowledge that can be used so effectively to destroy can be used also to build.

FRUITS OF THE UNION

Even in peace, man is constantly at war. There is the urgent and ceaseless war against disease, both physical and mental; the war against the insect world; the war against the forces of nature; the battle against hunger. We fight constantly against our social sicknesses of racial discrimination and disrespect for law and order, and against ignorance and the abuse of power. Our success in the constant warfare of peace depends, in my opinion, primarily upon how well we utilize our scientific resources—how well the partnership of knowledge and man, or science and politics, is joined in the battle.

The men of science and the knowledge gained through their researches have not been partners or participants in the art of government. That art traditionally, and quite properly, belongs to the politician. It should continue to belong to the politicians, for they are essential in any government, but they should be strengthened by the scientists and the knowledge gained by their methods.

The scientist has chosen to be a specialist. His work is so demanding that he cannot assume formal political and social responsiblities. The demands of the laboratory and the nature of his work have made the scientist somewhat inaccessible. Furthermore, there has been a lack of understanding of scientists and their work, as well as a patronizing attitude toward men of science on the part of many men of affairs, an attitude that has prevented the union of these two groups and thus has prevented our society from receiving the full benefit of their joint talents.

The achievements of science under the awful pressure of war have increased the public awareness of the resources of science and its enormous potentiality for the betterment of our lives. The explosion over Hiroshima blasted everyone—scientists, politicians, and plain citizens—into thinking about science and its implications for the future of the world, not only in terms of better things, but also in terms of a better political, social, and economic world.

During the past year, no citizen has thought harder or worried more about the implications of science and technology than has the scientist himself. He, therefore, emerged from the laboratory into the political world to try to explain to the politician what science is, its conditions for healthy growth, and how it should be supported and utilized for public good rather than for the destruction of our world.

The task of explaining their views to the political public proved to be arduous and exasperating, but enlightening also, to the many scientists who took the plunge. The effectiveness of the scientist's attempt to influence public opinion and thinking cannot yet be assessed. It certainly was less effective than he hoped; however real gains were made in the formulation of national policy. One thing is clear, however—the scientist must continue to tell the public what he thinks about both national and international problems facing the nation. During the 1945–46 Congressional sessions, we saw the United States Congress consider and debate two major bills relating to federal support and control of scientific research; the Kilgore–Magnuson Bill establishing a National Science Foundation, and the McMahon Bill establishing federal control of atomic energy. The latter became law, and the former was passed over without action. In addition, at least five other bills concerning government support of science were thrown into the hopper; some for political purposes, some for apparently no reason at all. Even Mrs. Clare Boothe Luce tossed in a bill.

Behind the scenes in Washington, the political and legislative struggle went on for a year. It was a fruitful struggle, not in terms of actual results, but because it introduced the politician and the scientist to each other's brand of thinking, to each other's problems, with each striving to understand the other. The method used by the scientist in his thinking is a process of reasoning *from* facts to conclusions which fit all the facts. Such thinking often leads to new and unexpected conclusions. On the other hand, the politician reasons *from* conclusions into which he attempts to fit all the facts *if* he can. It takes inspired political thinking to produce a result which fits all the facts!

Although the United States Government has supported certain types of scientific research, particularly in the agricultural field, for many years, the issue as to how far it should go in the support of a truly broad and national scientific program still remains unsettled. Fundamental research, upon which all applied research and engineering depends, has been left largely to individual enterprise on the part of men of genius who are wise enough to know the long-term value of knowledge. The question now is, "Can scientific work be left to happenstance, to the vagaries of a vanishing philanthropy, to universities with slim research budgets, to an industry demanding immediate returns on any investment?"

The answer is, "No," provided that a way can be found for the government to support science without restricting

the freedom of investigation and the exchange of knowledge, a freedom without which science cannot thrive.

A NATIONAL SCIENCE PROGRAM

There are several compelling reasons why our government must subsidize and support a national science program and not leave the matter to chance.

Our democratic system of government and our belief in private enterprise are grounded in our happy experiences with them, but that experience is only 150 years old. Throughout this brief span we have been operating in an expanding economy with new frontiers of land, increasing population, and new resources constantly opening for us. These frontiers, as we once knew them, now have vanished and our population is rapidly approaching a state of equilibrium. The wise man, who loves his way of life, asks whether his system can endure and flourish under the changed conditions he sees ahead.

While frontiers in the pioneer sense of the word are gone, the new and truly endless frontier of science always will be ahead. It contains vast riches for those wise enough to find and to train the explorers, and to equip the necessary expeditions. I have no doubt that science can guarantee its part in the development of a new golden age for our people, but I doubt and fear lest the people of the world and their leaders are not wise enough to govern themselves.

We are living in a world of cruel reality and the issue of national security is constantly with us. National defense no longer is a matter of the size of the Army, Navy, and Air Force. It is a matter directly related to the total strength and resources of the nation, of which the scientific laboratories are as vital as any other part. That nation whose scientific stature and health are the greatest is thereby the strongest and most secure. Finally, it is true, almost without exception, that the public eventually receives the benefits of any new knowledge. The public, therefore, has a direct interest in supporting research, just as it supports education on all levels, for the public is the final recipient. Both industry and the universities will continue to support scientific research—the former because it pays dividends, and the latter because they have been given the function of training scientists. Neither industrial laboratories nor university laboratories nor federal laboratories can do the whole job. Each has its place in the total scientific picture. There are vast areas of research in nutrition, psychiatry, psychology, chemistry, physics, sociology, and economics, of such a long-range nature and of such complexity that only the government can make the required investment, the returns from which will come far in the future.

The establishment of a sound federal policy for the promotion of science is a matter of direct concern to everyone here today, and each of you must do your own thinking about it—more than that, you must act after you have thought.

Any federal policy or any legislation that may come in the future must result from the scientist and the politician working together. It must satisfy the scientists' demand for high scientific standards, not only in theory, but in practice. It must be sufficiently inclusive not to leave any essential group unrepresented. It must contain provisions acceptable to the legislative and executive branches of the government, since all legislation must be so designed as to work in the political sense. To make our Congress adopt a fruitful policy and pass good legislation depends upon successful political methods. All legislators must be informed regarding the merit to be sought and the evils to be avoided. Any policy or legislation must have a vast majority of scientists and educators supporting it. Legislators must be convinced that the federal support of science is worth the cost. The general public must be sympathetic; at least not unsympathetic. The art of government has its own methods of accomplishing its work. The scientist can do only what any other citizen can do with political methods—study them and follow them until the job is done.

I close with a warning. When our government, or any other government, supports something, it exacts a measure of control. There is no such thing as controlled science, for when science is controlled it no longer is science. We have just seen a nation of great and talented people destroyed in Germany, which, under Hitler, abolished intellectual freedom and passed the control of scientific thinking into the hands of a government of venal and foolish politicians. It can happen here—unless all scientists become citizen-politicians. Making science and politics work together for the national welfare is your job now.

Sliding Mercury Contact. It always has been difficult to transfer high currents from a stationary to a moving object. Machinery requiring large currents has used heavy flexible leads, flexible copper laminations, or large numbers of carbon brushes, and has had a limited length of motion for moving current-carrying parts. A high current capacity sliding contact developed to improve these conditions is described by M. B. Austin of Coatesville, Pa., in the *Iron and Steel Engineer*, September 1946. It consists of a copper container holding a mercury bath in which is suspended, so as to be separated from the container, an upper copper contact, the contact surface of which is covered completely by mercury. Either the container, upper contact, or both may move, motion between them causing both contacts to slide on a mercury film one eighth of an inch thick. Temperature rise in this type of contact is 15 degrees Fahrenheit at a current density of 500 amperes per square inch per 1/64 inch length of current path through the mercury, and 24 degrees Fahrenheit at a current density of 750 amperes per square inch under the same condition, both rises occurring during a 5-minute period of contact operation.

Slow-Acting Relays

I. Historical Aspects

ARTHUR BESSEY SMITH
FELLOW AIEE

Slow-acting relays have been the unobtrusive little behind-the-scenes timing-control elements in various kinds of electrical installations for so long that they now are taken for granted. Although supervisory control, telemetering, and protective relay systems for electric power systems, have been covered extensively in technical literature, the role played by the slow-acting relay for the most part has been passed over. In a like manner, much has been written about machine switching in automatic telephony, but little has been said about the slow-acting relays at the heart of the works. Aware of this deficiency in published information, the AIEE committee on communications, under direct sponsorship of the subcommittee on quick and slow-acting relays, arranged a conference at the 1946 AIEE winter convention for the purpose of preliminary exploration of the subject and of the interest in it. The essential substance of that material is presented in these articles.

IT IS NECESSARY to understand clearly what is meant by a relay which is slow to operate or slow to release. It is not a matter of definite time, such as so many milliseconds. It is relative to the speed of the great mass of relays used in the particular field under consideration. In a field where most relays operate in 5 or 10 milliseconds, one which intentionally takes as much as 30 milliseconds to operate is clearly a slow-operating relay. But in a field in which the great run of relays take 40 to 70 milliseconds to operate, a relay would need to delay its operation as much as 150 to 200 milliseconds to classify as slow to operate.

Today slow relays are used in a number of fields, such as electric power, telegraph, and telephone. Most of these fields began to use slow relays in the era from about 1900 to 1910.

ELECTRIC POWER

At least as early as 1904 there was in use a relay of the solenoid and plunger type, fitted with a leather bellows having a small hole in one head. The slow escape of the air delayed the completion of the stroke of the plunger to the desired amount. Some relays of this type are still in use.

Later relays of the induction motor type were used, in which the delay was obtained by means of a disk. Inertia played a part in the delay and sometimes small permanent magnets were added, to cause eddy currents in the disk, similar to the structure in the integrating watt-hour meter.

TELEGRAPH

Hand-operated Morse telegraph sets seemed to need no slow relays. The circuit must be under just as perfect control as is practical, so that the operator can make dots, spaces, and dashes, long and short, at will.

Before 1909 a ticker system used by the New York Quotation Company employed a "time relay." It had a long stroke to obtain a desired delay in operation.

In the same era there was a step-by-step telegraph printer which had a "press magnet" whose armature was heavy enough to furnish considerable inertia. During the flow of rapidly alternating impulses the armature could not respond. But on steady direct current it operated and printed the character. The same device was used in the Gold stock ticker and in the Essick page and line printer.

TELEPHONE

It was in the telephone field that relays fully came to their own, as to numbers and importance to the art. Because such large numbers of relays are needed, they must be relatively cheap. Because of limited space they cannot have the desirable cube form, but must be relatively long and slender. This enables the maintainer to get at any relay at either end without taking it off the shelf or rack. Early manual switchboards had no need for slow relays. In fact, speed was not considered.

The first automatic switches were operated by the hand of the subscriber, using push buttons or strap keys, with so many pushes for each digit. Time was not much of a consideration, for the two wires of the subscriber line were grounded in turn in accordance with a definite code. Numerical impulses were sent over one line wire and one impulse over the other wire operated the "numerical separator" so that the next group of impulses on the first wire would operate a different magnet. There was no need for slow relays.

The entrance of a definitely slow-to-release relay is interesting.

In 1900 the New Bedford, Mass., automatic telephone exchange was installed. Each selector had two series line relays, each 30 ohms in resistance. These remained in series with the line wires after the connection was extended to the next switch. Thus the completed con-

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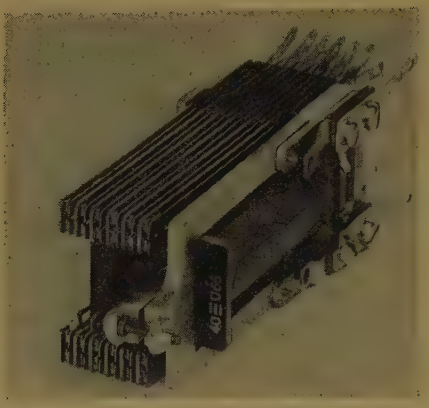

Figure 1. The *U*-type relay

nection had several pairs of 30-ohm relays in series, through which the subscribers must talk (local battery). To improve the voice transmission, each relay was shunted with a 150-ohm noninductive resistance.

In 1901 the Fall River, Mass., exchange was installed. It had the same 30-ohm series relays in each switch, but the iron core of each relay was covered by a copper tube, whose wall was about 0.80 millimeter thick. The same copper sleeve was applied to several magnets used in lifting and rotating the switch-wiper shaft. In the latter case it materially reduced the spark produced at the contact of the relay which controlled such magnets.

In the same plant, and at about the same time, it was discovered that short-circuited turns in a relay winding, as well as the copper sleeve on the core, had a strong delaying effect on the release of the relay. In fact, it was found that if the copper sleeve were thick enough, it tended to run the impulses together.

At least as early as 1904 there were a number of a-c relays in use. The armature, instead of being pivoted at one end, was pivoted somewhere near the middle, and it was capable of rotating as much as 40 or 45 degrees before closing the contact spring. This had in it great possibilities as a slow d-c relay, though it seems not to have been used that way.

In 1904, during the month of July, an engineer of the Automatic Electric Company devised the first 2-wire common battery subscriber line circuit, combining quick and slow relays in numerical selection. At the subscriber station there was no ground connection (except for the lightning arrester). When the subscriber seized the line by closing the loop, a quick line relay operated and pulled up a slow relay. The latter grounded the third wire and prepared local circuits for receiving numerical impulses. As the subscriber dialed the first digit, the line was opened several times, each opening causing the operation of the motor magnet which moved the switch wipers. During this time the slow relay remained operated. A second slow relay was associated with the motor magnet—this relay operated and remained operated until the end of the group of impulses. Then it released and prepared the local circuits for the next group of impulses (the second digit of the call number). These slow-to-release relays were made so by copper sleeves.

In 1908 the Kellogg Switchboard and Supply Company needed a trunk circuit between an automatic exchange and one of their own manual switchboards. The company devised a circuit for the manual end of the trunk line which would ground both the trunk wires, then clear both from earth. This was done by the use of two quick relays and one slow-release relay. When the operator pulled down the connection, the first quick relay let go and connected the two trunk wires to earth, also through the other quick relay. The latter operated on the release current, and cut off the slow relay, which soon released and cleared the lines.

By 1910 the slow relay had been changed by its use of a copper collar instead of the copper sleeve. By placing the copper collar on either end of the relay core, great differences in performance can be obtained. Because of the slender shape of the electromagnet, if the collar is at the armature end of the core, the delay in operating time may be as much as 75 to 100 milliseconds. If it is at the "heel" end of the core, the delay to operate may be only 10 to 15 milliseconds. This difference is a result of the magnetic leakage between core and heel-piece, which would not be very effective if the magnet had a cubic shape.

In 1930 there was a return to the sleeve for the release and magnet relays, in order to diminish the cost. The cost of the copper and its labor was reduced about 40 per cent and that of the complete coil from 10 to 15 per cent. There has been no impairment in the performance, but the copper collar is retained wherever its specific performance is of value.

Other means, such as weighted springs or clockworks with air for governors, may be used for slowing the relay but only the copper collar and sleeve seem to have obtained any significant degree of usage in the telephone field.

II. Use in Telephone Switching

H. N. WAGAR

A TYPICAL RELAY, used by the millions in telephone offices, in shown in Figure 1—the *U*-type relay. This particular type is not a slow relay; its action time is about 1/20th second or less. It can be slowed, however, to give times in the range from 50 to 500 milliseconds, which is the ordinary range of delay in telephone relays. These delays can be obtained—at reasonable cost, with reasonable accuracy, and without too great a sacrifice in number of contacts—through application of the principles to be outlined. Primarily,

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these delays are obtained by the use of short-circuited turns to oppose a change in magnetization, and by using the properties of the relay upon disconnection of the circuit, not on closure, because of the greater stability of a relay upon its release. The subject of this article then narrows down to the principles underlying slow-release relays.

At the bottom of Figure 2 appears the "slow-release formula," by which almost any slow-release problem may be explained. Because of its importance, the steps by which this formula is reached will be outlined. At the top of the figure is a simplified relay structure. When the switch S is closed, the armature is attracted from its unoperated position to one against the core. The condition of magnetization is represented by the dotted φ versus Ni curve at the right, the point corresponding to φ_1 representing conditions after the armature has come to rest against the core. In this operated position, the armature experiences a mechanical force, F, resulting from the deflected springs trying to restore it to normal, and a magnetic force, P, holding it against the core. This magnetic force depends on the flux in the air gap, and the pole face area, S, according to the formula

$$P=\frac{\varphi^2}{8\pi 980S}\text{(in grams)}$$

When the switch is opened, the applied ampere-turns eventually will drop to zero, and the flux follows according to its demagnetization characteristic. With the flux, of course, the pull also falls off.

The release time, therefore, will be the time, after disconnecting the switch, at which equilibrium between magnetic and mechanical forces is just unbalanced. It is, then, the time for the flux to fall from its steady value to the value of flux at which the pull, P, resulting from magnetic attraction equals the spring force, F. Looking at the heavy curve to the right, it depends on the time for the flux to travel its downward φ-Ni characteristic (its demagnetization curve) to the just-releasing point. This time would be negligible were it not for the eddy-currents induced in the second winding shown, a short-circuited winding which opposes any change in flux

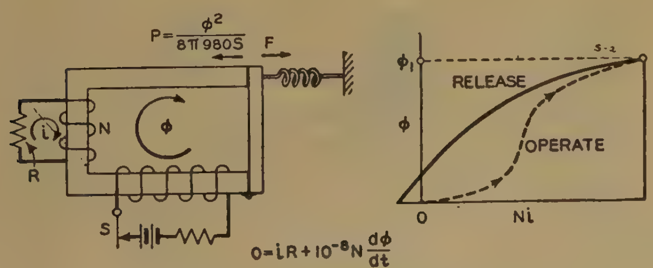

Figure 2. Slow release relay

Slow release formula:

$$t=10^{-8}\frac{N^2}{R}\int^{\varphi_1}\frac{d\varphi}{Ni}\text{ second}$$

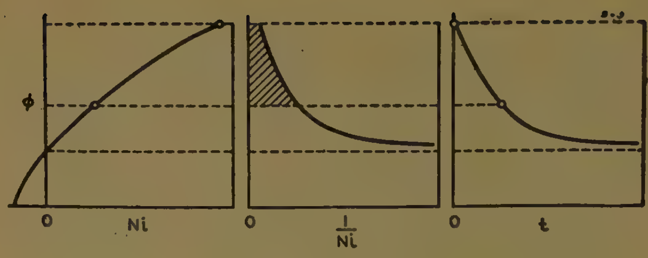

Figure 3. Slow release formula split into conductive and inductive terms

Slow release formula = (*conductive term*) × (*inductive term*)

linking its turns. All this may be expressed mathematically by writing down the voltages around the circuit: the IR drop plus the back electromotive force, which results from a changing flux, of $N(d\varphi/dt)$. This is shown written in correct units, and when terms are rearranged gives the slow-release formula already mentioned.

When this formula is used, it is convenient to split it into two separate terms, as shown in Figure 3. The N^2/R term has dimensions of a conductance and is called the "conductive term"; it depends solely on the dimensions of the coil and the quality of the conductor. The remainder of the expression is called the "inductive term":

$$10^{-8}\int^{\varphi_1}\frac{d\varphi}{Ni}$$

It is a function of the iron quality and the iron dimensions. In other words, it depends on the descending φ–Ni relation of the magnet, a relation amenable to both measurement and prediction. By replotting the flux against $1/Ni$, the information is arranged in the form $\int xdy$, which expresses the area included between the curve, the vertical axis, and the two ordinate limits. This area is shown shaded in Figure 3. The product of this area and the conductive term gives the desired time in seconds, varying as shown at the extreme right of the figure. A more detailed examination of each of these terms will be made next.

The manner in which the conductive term varies is seen in Figure 4, where the conductance of a coil (in megamhos per inch of length) is plotted against the ratio of outside to inside radii. The heavy curve, representing a solid sleeve, is seen to give the greatest delay in a given space, as might be expected. The other curves show the results for the extremes of typical windings where the value of e represents the "copper efficiency" of magnet wire; that is, the ratio of copper volume to total volume.

The inductive term in Figure 5 can be used to estimate effects of iron, gap, or other dimensions. As an illustration, the three curves may be considered as those for three different operated gap-lengths, differing only by a few tenths of a mil from each other. The replot of these values against $1/Ni$ shows how the smallest gap

gives the largest area, and consequently the longest time, for a given spring force. The dimensions considered here are in the order of the thickness of the finish on the parts, so that the important effect on time resulting from wear of the finish or manufacturing variations can be recognized.

The three cases being compared here are seen to differ widely in time upon the assumption that all are adjusted to the same operated load. Should it be possible to change the loads so that in each of the three cases the just-release ampere turns were the same, it will be evident by reading across from the set of curves on the left side of Figure 5 to that on the right, that the releasing time will be approximately the same for each case. This relationship between just-release ampere turns and time is the basis for large-scale manufacture of the slow-release telephone relay known as the *Y*-type relay to be described further in the following. So long as springs are provided to give the needed releasing load, the time is known when the release *Ni* are known.

The curves also could be used to analyze the effectiveness of various pole face areas. In general, a large pole face will be found best suited to long delays, but an optimum dimension sometimes will be indicated by this type of analysis. The use of these methods gives the designer reliable results. Figure 6 shows agreement between prediction and experiment in a particular test sample; where the heavy curve shows the calculated time plotted against releasing *Ni*, and the circles show measured points.

To summarize, then, the design features of the slow release relays are as follows:

1. Short-circuited windings.
2. Low-reluctance gaps.
3. Low-leakage magnetic design.
4. Saturated iron (high current).
5. Low spring forces.
6. Massive moving parts.

The designer will build his slow relay using a solid sleeve of copper wherever possible.

The smallest possible operated gaps consistent with corrosion and wear resistance will be used, and a large pole face area will be helpful. An efficient magnetic circuit, operated into the saturated region, will be best. Light restoring forces will require a longer flux decay time to reach. For completeness, massiveness of the moving parts is mentioned, though it usually is of secondary importance.

Such principles are embodied in the *Y*-type relay shown in Figure 7. This relay uses copper sleeves, iron saturated by 200 to 250 ampere turns, a low-leakage magnetic circuit, nickel finish for corrosion protection, and a thin chromium finish for wear protection. While it uses nearly iron-to-iron contact in the operated position, an important feature is a slight sacrifice in operated gap reluctance to obtain greater stability—the spherical gap surface (visible in the figure), which gives uniformity at the gap in spite of misalignments encountered with normal manufacturing practice. Its other feature, as previously mentioned, is its ability to be adjusted for a given time requirement simply by adjusting to release on a specified value of ampere turns. The extra springs provided to insure that the releasing force can be attained in all cases will be seen in the figure.

III. Telephone-Type Relays

ANDREW W. VINCENT

THE TIME INTERVAL required for the closing or opening of a relay contact after the current is applied or removed, is dependent upon the balance of mechanical forces acting on the mass of the armature. These balanced forces are the spring load force, the magnetic operating force, the frictional force, and the force of acceleration. Because changes in the magnetic forces occur when the armature is unoperated, while it is moving through its stroke, and also after it is operated, it is necessary to show the relationship of the magnetic forces as a function of armature stroke at constant ampere-turn values and the magnetic forces as a func-

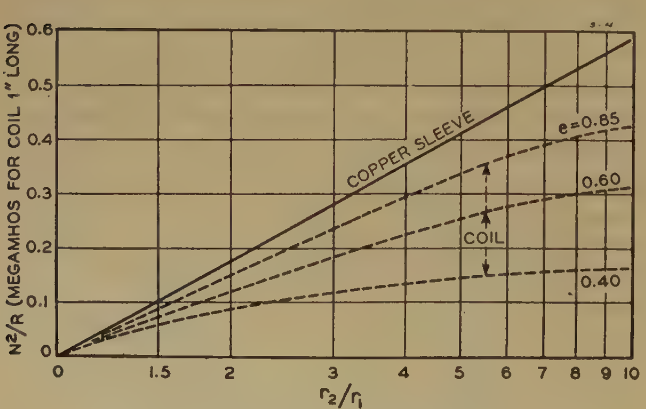

Figure 4. The conductive term N^2/R

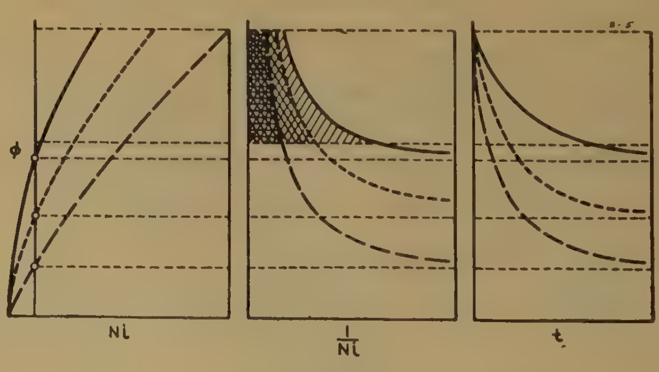

Figure 5. The inductive term $10^{-8}\int_{\varphi}^{\varphi_1} \frac{d\varphi}{Ni}$

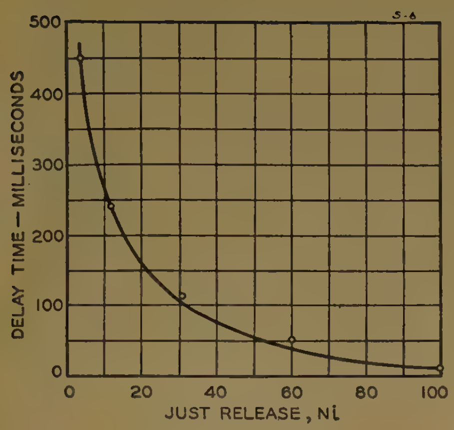

Figure 6. Agreement between prediction and experiment in a particular test sample

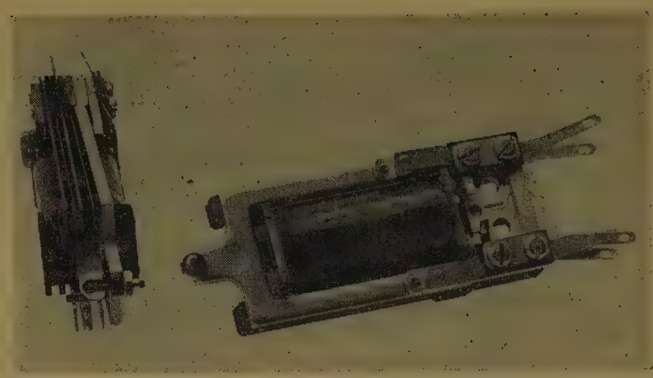

Figure 7. The *Y*-type relay

tion of ampere turns at various armature stroke positions. The curve of Figure 8 shows these relationships in a typical telephone-type relay structure carrying an operating load of 12 make-contacts and operating with zero residual air gap.

The left half of the curve sheet shows the magnetic force versus armature stroke at different ampere-turn values, and the spring load force versus stroke of the 12 make-contact spring load. The dotted curves show the build-up and build-down forces under typical operation. The right half of the curve sheet shows the build-up of force at operated and unoperated stroke positions after demagnetization and build-down after 600 ampere-turn saturation. Starting at 0 and 0′, and progressing through the numbered points back to zero, shows the static forces throughout the armature stroke. The following is a detailed discussion of the cycle.

From 0′ to 1′ represents the build-up of the magnetic force in the unoperated armature position to the point where it is greater than the spring load force at 1, at which time armature motion begins. The time required for this part of the cycle is dependent upon the time required for the current in the coil to build up to the required value as given by

$$i=(E/R)(1-\epsilon^{-Rt/L})$$

where L equals the inductance of the coil with the armature in unoperated position.

When the eddy current losses in the iron are large, the accompanying counter flux may cause the time to be much longer than the foregoing equation would indicate. Advantage has been taken of this in adding short-circuited turns to the winding to increase the operating time. The short circuit may take the form of a relay winding or a solid copper collar. The location of the short-circuiting winding must be kept close to the armature to prevent the leakage flux from acting on the armature. Inasmuch as the armature is open, there is more of a tendency for the flux to take the leakage path.

Points 1 to 2 indicate the rise of force as the armature closes. Although the spring load would be lifted by 110 ampere turns as indicated by the fact that the spring load is less than the 110 ampere-turn force curve throughout the stroke, it is necessary to saturate the structure with 600 ampere turns in order to show the most severe release condition. When the armature has completed its stroke, the ampere turns have increased to 150 which is shown at 2. The contacts close at the point indicated, the elapsed time being that necessary for the changing force, represented by the difference between the spring load force and the dotted force curve, to move the effective mass of the armature and springs to the point where the contacts close. As the current has not reached its final state, the force increases at a constant closed gap from 2′ to 3′ which is outside the curve sheet. The points 2 and 2′ would not coincide exactly because of the different paths taken, but the difference is assumed to be negligible. Curve 0′2′3′ is similar to the normal magnetization curve of the iron, and the upper curve is similar to the upper part of the hysteresis loop, except that the force is a squared function of the flux. The equation is

$$F=\frac{\phi^2}{7{,}840\pi A}\text{grams}$$

when

ϕ = flux in maxwells
A = area of the gap in square centimeters

When the relay coil circuit is opened, the force decreases along the curve 3′ to 4′ until the magnetic force is less than the spring load, at which time armature motion begins. The elapsed time is again dependent upon the effectiveness of the eddy current paths in building up currents resulting from the changing flux which builds up counter flux to oppose the change in flux and thus maintain the force. Inasmuch as the armature is closed, the position of the short-circuiting conductor is not as important as in the case of the unoperated armature position. The rate at which the circuit contact is opened also affects the time, especially if there are no short-circuited turns, since the large change in flux builds up high voltages which cause current to flow as an arc.

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The point 4′ indicates that the relay will release on 35 ampere turns.

The dotted release force curve 4 to 0 indicates the decrease in magnetic force as the armature opens. The force decreases rapidly as the armature is opened inasmuch as the flux path is broken. The time elapsed until the contacts open is dependent upon the applied forces and the mass of the system as in the operating case.

In practice, the determination of relay release and operating times is limited because of the large number of variables.

The value of the voltage applied to the coil is not an important factor in the consideration of release time because the magnetic structure is near saturation with the armature closed, under average operating conditions. The release time is largely dependent upon the rate at which the flux decays. The time required for armature motion can be neglected in slow releasing relays as it is a small part of the total release time. By knowing the time required for the force to drop to the operated spring load force at various residual air gaps and with a specified short-circuiting collar, the release time can be determined. With a large copper sleeve on the core release time as high as one-half second can be achieved without failure to release.

The decrease in height of the residual projection, resulting from operation wear, causes a greater increase in release time when smaller residual setting adjustment is used, inasmuch as the change resulting from wear is a greater percentage of the total residual air gap. When the effect of magnetic iron aging, which may double the residual force over a period of time, is taken into account along with the accompanying decrease in spring load force, it can be seen that the release time increases with the number of operations, approaching the point where the armature will fail to release.

In determining the operate time the applied voltage cannot be neglected. Over the normal operating ampere-turns range the introduction of the unoperated armature air gap into the magnetic circuit shifts the operating range of the iron to a lower region such that current changes are not masked by saturation. For this reason the applied voltage must be considered inasmuch as it, with the L/R ratio of the coil, determines the current in the coil as a function of time. The inductance of the coil is less with the unoperated armature air gap which is responsible for a maximum practical limit which appears to be about one-tenth second operating time delay.

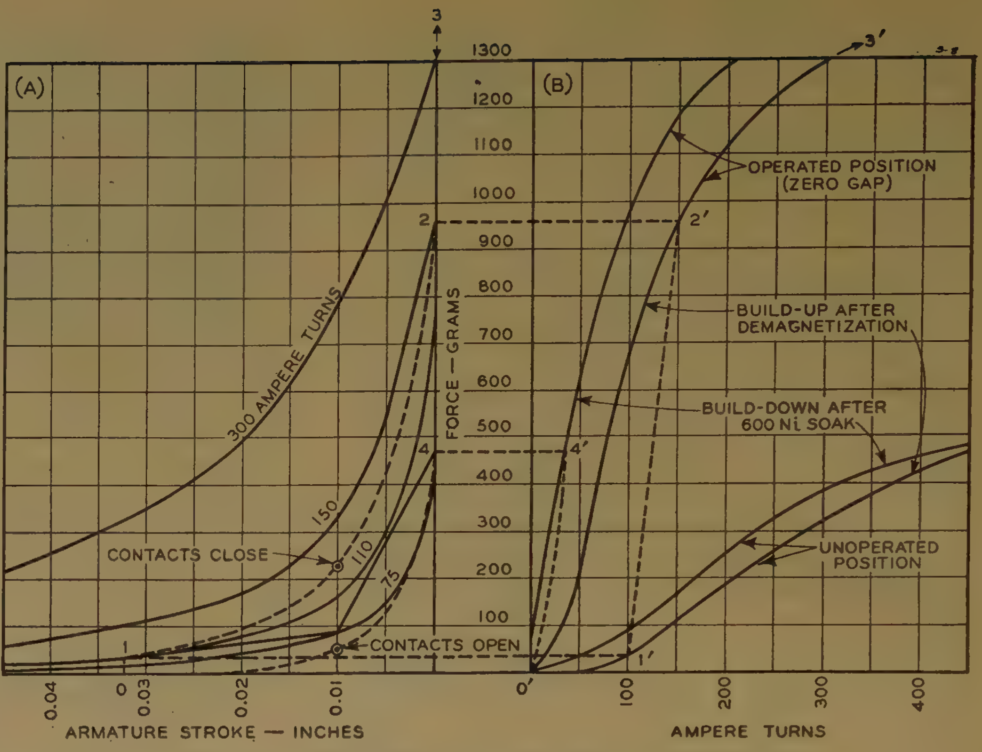

Figure 8. Relationship of magnetic forces as a function of:

A—Force versus stroke at constant ampere turns B—Force versus ampere turns at constant stroke

IV. Use for Power Purposes

E. E. GEORGE
FELLOW AIEE

THE APPLICATION of telephone relays in the power field is not new. These relays have been in use for years by many utility companies on special applications. The larger power manufacturers have used them as elemental units in telemetering, selective calling on carrier telephone circuits, carrier relay protection, protective distance relaying, and also in supervisory control.

Extensive use has been made of standard telephone relays in the regular, slow operate, and slow release types. In addition to the standard cylindrical-type relays the following relays or relay devices have been used by some power utilities:

1. Rotary selector switches.
2. Morkrum cable relays.
3. Double throw polarized relays.
4. Weighted reed timing relays.

E. E. George is electrical engineer, design and construction department, Ebasco Services, Inc., New York, N. Y.

5. Toll line ringing assemblies for 20-cycle ringing and for 135-cycle ringing.

The major use of such relays is as auxiliary devices, but they also are used a great deal for automatic reclosing, carrier signaling, private telephone line ringing, timing, sequence control, interlocking, and time delay purposes. The characteristics of standard relays are visualized easily if plotted on log-log paper.

The only major disadvantage of telephone relays is in their mounting requirements but this largely has been overcome by the design of a special front-of-board mounting case with studs for rear connection.

The advantages of these relays are their low price, small space requirements, wide range of contact assemblies, resistance range from $1^1/_2$ to 15,000 ohms, reasonably accurate calibration, dependable design and test procedure, adequate engineering data, good delivery (normally), and low energy requirements.

Organization of the Engineering Profession

ELGIN B. ROBERTSON
FELLOW AIEE

This searching consideration of the question of organizing or reorganizing the engineering profession is presented for the purpose of stimulating additional thought and discussion by the entire AIEE membership. The author, speaking in the light of experience as president of the Texas Society of Professional Engineers and vice-chairman of the AIEE membership committee, supports Plan B of the various proposals of the AIEE planning and co-ordination committee's professional activities subcommittee (EE, Apr '46, pp 169–73). He urges that collaboration and co-operation are more effective than attempted consolidation in view of the great diversity of interests among engineers.

IN CONSIDERING the organization of the engineering profession, there are two distinct phases which must receive attention—the *technical* and the *professional*.

The technical side of our profession is well organized. We have the Founder Societies and a host of other societies, each devoting itself principally to the dissemination of knowledge to its members. Our AIEE is more or less typical of these technical societies. It is certainly a going concern. It is the largest of this group. It receives a continuing infusion of new ideas through the new members, and the new local and national officers elected periodically. Certainly, as an organization, AIEE does not need any particular "going over." Any radical change in its organization, or of its set up, would not seem to be warranted, and I doubt that such changes would meet the approval of our membership. All these characteristics are but indications of a successful organization.

Let us not lose sight of the fact that our AIEE and other major technical societies are, as a rule, old organizations representing a wealth of accumulated experience and well-earned prestige. Let us remember that the success they now enjoy is the result of the thought, the time, the energy, and the money put into them by their members over a period of a half-century or more. Major changes in either status or major objectives should be approached with great care.

When we consider the professional side we find an entirely different situation, and it is with this phase that I wish to deal mainly.

What is a profession? One answer to this question can be gotten by referring to "Webster," but one will find only the definition of the *word*. Surely our engineering profession is more than can be defined by a single word, no matter how comprehensive that definition may be. But, regardless of any definition, I believe that a profession is what its members want it to be; inevitably, it is whatever they make it. If we want our engineering profession to be strong, virile, and honorable, then we, its members, must breathe into it the desired life, vitality, and honor through the integration of our individual actions. If we do not want this kind of a profession, then we can refuse to give to it our time, our energy, and our money. If we are

Based upon a talk given at the AIEE South West District meeting, San Antonio, Tex., April 16–18, 1946.

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personally selfish or thoughtless we may conduct ourselves so as to get the maximum benefit for ourselves, regardless of the effect such conduct may have on the welfare of the whole group. Some of us are doing just this, although the vast majority are not.

Every engineer has his technical needs and his professional needs. Both of these concern his welfare and so he should devote himself to both of these activities.

I have been told time and time again that engineers cannot divide their loyalty between technical and professional societies and that they do not have enough spare time to devote themselves to both. I have been associated with AIEE for approximately 30 years. I started out as a Student Member and have advanced through the various grades to that of Fellow. I have been a Section secretary and Section chairman, have served on numerous committees, and as a vice-chairman of the membership committee for the South West District. I have belonged to the Texas Society of Professional Engineers, and the National Society of Professional Engineers for several years. I have been a chapter secretary, a chapter president, a state director, and in January of this year, I completed my term of office as president of the Texas Society of Professional Engineers. Has this activity in the professional society lessened my interest in the AIEE? By no means! As a matter of fact, it has enhanced my interest in and appreciation of AIEE. This division of activity is a manifestation of breadth rather than division of professional interest. You cannot divide your loyalty to yourself; nor can you give to your societies (technical or professional) *only* time that you consider to be *spare* time if you wish them to prosper and to achieve the results you desire.

A technical society such as the AIEE cannot be an all-inclusive society. The technical problems of the electrical engineers are not the same as the problems of the civil, petroleum, or other engineers. However, a professional society can be all-inclusive because the professional problems of the electrical engineers are identical with those of all other engineers. These professional problems are not limited to the engineers in private practice but are the concern of all engineers.

Why are we interested in organizing the engineering profession and why are we discussing these professional problems? Actually, if we will be honest, we must admit that this interest is aroused by our desire to have the public give us the recognition, as a profession, to which we think we are entitled.

All right—we admit that we want public recognition and esteem as a profession. How will we get it?

Who is an engineer? Answer this question by taking a piece of paper and writing down your own definition of an engineer. If you will take the definitions given by ten different engineers, you will have ten different answers.

Who is a doctor? He is one who is licensed to practice medicine by the laws of his state. A doctor of medicine degree from a medical school, or years of experience, do not entitle him to practice. It is the license issued by his state that is the mark of identification and authorization. Who is a lawyer? He is one who is licensed to practice law by the laws of his state.

By the same token, I believe that we should define an engineer as one who is licensed to practice engineering by the laws of his state. If we of the profession cannot put a label on the engineer, then how can we expect the public to recognize an engineer when it sees one or hears one? We must be able to point definitely to a particular individual and say: "This is an engineer." I believe that the legal license is the best answer to this simple identification.

I believe that most of us will admit that the lawyers have influence in their communities. I believe that we will concede further that the public is conscious of the professional status of the doctor. Would either of these be in their present position without the Bar Association and the Medical Society? A professional engineering society is essential to professional recognition and standing.

The professional activities subcommittee of the AIEE committee on planning and co-ordination has made a report on four plans for organizing the engineering profession (*EE, Apr '46, pp 169–73*). This report is worthy of every engineer's thought and consideration. I believe that the course presented by "Plan *B*" of that report is the best to follow. Plan *B* leaves the AIEE and other technical societies to carry on as they now do and calls for a professional society to carry on all activities that are of a nontechnical nature. I believe that Plan *B* will do the job we want done.

As the all-inclusive professional society called for in Plan *B*, I personally favor the National Society of Professional Engineers because

1. It is already in existence as a going concern and ready to serve.

2. A license is a prerequisite to its membership.

3. It is a thoroughly democratic organization. Its members are elected locally. The state societies and the national society cannot elect a member. A state society is primarily a co-ordinating agency of the local chapters in that state and the same is true of the national society in its relation to the state societies.

This professional problem is primarily of local concern and since all power in this organization originates with the individual member and his local chapter, it is particularly effective in handling the questions that arise. The local chapters can work through the state and national societies in taking care of those problems that are more than local in scope.

This organization frankly says that it is concerned with the welfare of the individual engineer, and since this is the main concern of all I think it is worthy of the support of all engineers who want an engineering profession that is entitled to the esteem and respect of the public.

Great Lakes District Meeting

Conference Papers Digested

Conferences on subjects varying from electric machinery to basic sciences were held at the Great Lakes District meeting of the Institute held at Indianapolis, Ind., from October 9–11. The majority of the papers presented were of the conference or District type and will not be published formally. The following material summarizes the papers presented at the meeting.

CONFERENCE ON ELECTRIC MACHINERY

The conference on electric machinery was presided over by Chairman M. S. Oldacre (M '42) Commonwealth Edison Company, Chicago, Ill. The conference covered a wide range of material from theoretical design to present-day commercial standardization of small motors. Outstanding features of the meeting were the papers on magnet wire and fractional-horsepower motor standardization. The paper by C. P. Potter on standardization stimulated the greatest amount of discussion, and it seemed clear that adoption of these standards will be of great benefit to both users and manufacturers in this $100,000,000 a year industry. (*EE, Nov '46, p 451*)

"Asynchronous and Single Phase Operation of Synchronous Machines" by A. W. Rankin (M '45) General Electric Company, Schenectady, N. Y. This is the fourth in a series of analytical studies of synchronous machines made by Mr. Rankin, the other three already having been published in the 1945 AIEE *Transactions*. In this current paper, Mr. Rankin presents an analysis of the asynchronous and single phase operation of a synchronous machine based upon the generalized equivalent circuit of Kron. The fundamental equations of Park are used as the starting point, but the succeeding mathematical operations are guided by the tensor publications of Kron. The asynchronous torque and current relations are presented in terms of the currents and voltages of simple equivalent circuits; also, a numerical comparison is given of the asynchronous and single phase operating characteristics of synchronous machines with complete and with incomplete end rings on the damper windings. This comparison, the author states, has direct value in design engineering, not only as a study of the various torques, but also as a picture of the distribution of the damper-bar currents in the leading and the trailing pole halves. The methods presented are intended for use with the a-c network analyzer, and all torque and current expressions are given directly in terms of current and voltage readings obtained from the network analyzer. Equivalent circuits for constant-speed asynchronous operation are derived, and numerical examples of the application of these equivalent circuits are given. Equivalent circuits for constant-speed single-phase operation also are derived, and a single numerical example of the application of these equivalent circuits is given.

"Fractional-Horsepower Motors for Operation on Thyratrons" by W. R. Goss and F. T. Carlson (nonmembers) of the General Electric Company, Fort Wayne, Ind. The many special applications of motors to welding-head controls and other such equipment where the motors are operated by power from thyratron rectifiers have led to the development of "packaged" power rectifier equipment by several manufacturers for use in driving variable speed motors on many types of application. The availability of this packaged thyratron-rectifier power unit has increased the tendency to utilize standard motors, a practice which the authors report as being not entirely satisfactory because of the additional heating, poor commutation, and noise, which results from the use of thyratron-rectified power on motors not designed for that kind of a power source. Specially designed and specially rated motors have been developed for use with thyratron-rectified power, thus making the variable speed qualities of a d-c motor fully and readily obtainable wherever a-c power is available, through thyratron rectification.

"Properties of Magnet Wire" by H. A. Smith (A '46) and W. G. Skinner (nonmember) General Electric Company, Fort Wayne, Ind. Magnet wire is one of the most important materials used in the construction of electric apparatus. More than 200 million pounds of copper magnet wire are produced annually in the United States. Although it is not generally realized, the magnet-wire industry is a precision industry. Tolerances which would be close in a precision tool room are not uncommon on small sizes of wire, such as the 0.0001 inch plus or minus that is allowed for wire of the size of an ordinary hair, about 0.003 inch. In the past, the acceptability of magnet wire was determined by trial-and-error methods, the design engineer winding such wire into a model of his apparatus and then testing it to destruction. Now, test methods for the evaluation of magnet wire are used by the wire engineer. An attempt is made to correlate the properties of insulated magnet wire—physical, chemical, and electrical—measured by wire tests to the properties desired in the finished electric apparatus. Of the 200 million pounds of magnet wire used by the electrical industry in a normal year, some 70 per cent is film coated—that is, enameled. The two main types of wire enamel used today are synthetic resin enamels and oleoresinous enamels based upon natural resins in combination with drying oils. Of the several types of synthetic-resin wire enamels now in use, the major one is enamel based upon polyvinyl acetal resin, the properties of which are far superior to those of a natural resin enamel for film-coated wire.

"New NEMA Fractional-Horsepower Motor Standards" by C. P. Potter (F '29) Wagner Electric Corporation, St. Louis, Mo. There are millions of fractional-horsepower motors in operation in the United States at the present time driving washing machines, stokers, oil burners, refrigerators, domestic water pumps, gasoline dispensing pumps and an almost limitless list of other devices including business machines. The public has become accustomed to having these devices operate essentially night and day with very few interruptions. Every fractional-horsepower motor has a horsepower rating stamped on its name plate, and this would seem to be a definite indication of its capabilities. However, this is not necessarily true, because motor characteristics frequently are modified to suit the desires of the manufacturer of the devices which they operate; some modifications being logical, others illogical. These and related practices have resulted in many fractional-horsepower motors not having the life expectancy which they would have if they were more consistently and conservatively designed and applied. To promote a more logical design and application of fractional-horsepower motors, the better to serve the interests of all parties concerned, the National Electric Manufacturers Association recently has adopted two important new fractional-horsepower motor standards: "Definition of Motor Output" and "Service Factor." These prescribed standardized readings, adherence to which would assure to the user a long and dependable service life, would enable the application engineer and the appliance manufacturer to select a dependably standardized motor thus obviating the need to test each application.

CONFERENCE ON ELECTRONICS

The conference on electronics met under the direction of O. W. Livingston (M '43), General Electric Company, Schenectady, N. Y. In summarizing, Livingston stated that a method should be developed by which all vacuum tube characteristics could be expressed in terms of volts, amperes, seconds, and temperature thereby simplifying description and application through standardization. It was interesting to note that the papers on electronics in measurements and aeronautical electronic devices concurred in the necessity for components and vacuum tubes possessing simplicity and ruggedness required in these applications.

"Electronics in Measurements" by R. J. Kryter (A '35) Esterline-Angus Company, Indianapolis, Ind. In spite of the fact that many types of measurements otherwise im-

possible can be made with the aid of appropriate electronic equipment, electronic methods admittedly have been slow in penetrating the field of instrumentation. Prejudice and sheer human inertia of course have contributed to the slowness of acceptance of electronic instrumentation, but it must be admitted also that it was logical for electronic instruments and measurements methods to be required to demonstrate both their facility and their dependability, and it must be admitted further that there is much room for refinement and improvements in ruggedness and simplicity of the electronic devices used in measurements, particularly in the vacuum tubes themselves. Electronic measurement methods must compete with existing mechanical, electrical, hydraulic, and pneumatic methods, and, initially at least, are most attractive in the fields not so readily served by the more established instrumentation equipment and methods. For example, electronic methods are most advantageously adapted to miscellaneous or special quantities such as light and sound intensities, color values, torque, gas composition, pH, or measurements under special conditions such as speed without physical contact. Similarly, electronic methods are indicated for the measurement of very small quantities, very rapidly changing quantities or conditions, and very short time intervals. Nonelectronic methods commonly require a human operator and manual manipulation, whereas the electronic method and equipment permits continuous and automatic measurement, and automatic control. Extreme sensitivity, great amplification, unlimited speed, smooth and easy control, possibility of automatic operation, are advantages. However, electronic instrumentation is not a cure-all; lack of ruggedness, difficulty of service and requirement for special service personnel, limited life, instability of calibration, and obsolescent components are challenging problems requiring and currently receiving attention.

"Frequency Performance of Thyratrons" by H. H. Wittenberg (A '46) Radio Corporation of America, Victor Division, Lancaster Pa. As there are many applications for thyratrons at frequencies above the standard power frequency of 60 cycles per second—such as in inverters, servomechanism controls, relaxation oscillators, radar modulators, and grid-controlled rectifiers for air-borne equipment, where the frequency range is from a few hundred to as much as 50,000 cycles per second—the performance of thyratrons at these frequencies becomes a matter of some interest and importance. Investigation reveals that at high audio frequencies the grid-controlled characteristic resolves itself into two characteristics—a starting characteristic and an extinguishing characteristic; the starting characteristic shifts negatively with increased frequency, and with increasing gas pressure; the extinguishing characteristics shift negatively with increasing frequencies, with increasing grid resistance, with increasing anode current, and with increasing gas pressure—the shifts also being a function of tube geometry.

"Microwave Magnetrons" by F. F. Reike (nonmember) Purdue University, Lafayette, Ind. In discussing the possible uses of magnetrons—as one of the members of the vacuum tube family—exclusive of radar, Doctor Reike pointed out that the cost of the tube (probably $1,000 each for small quantity production) and the design requirements for the dissipation of heat (about five per cent of rated power from the cathode) were two major problems with which to be contended. He described the steady power output possibilities as ranging from 0.1 to 5 kw, the pulse power or instantaneous output as ranging from 100 to 3,000 kw, with an efficiency of approximately 50 per cent. Aside from radar, the magnetron is considered to have possible applications in the communications field in connection with radio relay links, in industrial heating, and as a source of high power at high frequency for various research purposes. The pulse magnetron is not useful as a continuous-wave source.

"Electron- and Ion-Beam Instruments" by G. W. Dunlap (M '42) General Electric Company, Schenectady, N. Y. Modern beam-type instruments are dependent on a number of requirements common to nearly all. These are: an evacuated space; source of charged particles; means for projecting these particles; beam control; and means for utilizing beam. From the discovery of vacuum technique in 1650, the discovery of charged particles in 1853, and Edison's observation in 1883 of the famous "Edison Effect," the author traced the continuity of fundamental principle and technique in the successive development of various electron-beam and ion-beam devices ranging from the oscillograph to the mass spectrometer to the cyclotron–betatron–synchrotron family of accelerators.

"The Life of Electronic Devices Used by United States Army Air Forces" by A. P. Upton (nonmember) Minneapolis-Honeywell Company, Minneapolis, Minn. Stating that war experience showed conclusively that electron tubes are by far the weakest link in the entire chain of electronic control and other equipment, the author urged that designers and manufacturers take the steps necessary to make industrial electronic components "so rugged and so dependable that they can be soldered into the equipment circuits as easily and assuredly as one can now solder in a resistor or capacitor."

CONFERENCE ON POWER TRANSMISSION AND DISTRIBUTION

The paper on single-pole relaying and reclosing on a 132-kv system attracted considerable attention. It gave the first complete record over a number of years of a new development in tranmission engineering. Problems of rural electrification always seem to produce extended discussion, and this conference was no exception. Technical problems requiring solution relate to production of reliable low-cost devices for sectionalizing lines and to proper co-ordination of fuses and circuit reclosers. F. V. Smith (M '38) Sargent and Lundy, Engineers, Chicago, Ill. presided over the conference.

"Simplicity in Transformer Protection" by E. T. B. Gross (M '40) Illinois Institute of Technology, Chicago, Ill. This is essentially a review of the development and application of gas-operated relays as extensively used in Europe and in Great Britain for the protection of reactors and power transformers, an application for which differentially connected electromagnetic relays commonly are used in the United States. Whenever a fault in a transformer is beginning to form, heat is produced locally which leads to decomposition of insulating material with consequent production of gas. The nonelectrical relay is designed to be operated by the change in pressure within a transformer tank that is caused by the gas generated. Advantages of simplicity and sensitivity are claimed for the gas-operated relay.

"Rural Electrification in Iowa" by C. M. Stanley (M '40) Stanley Engineering Company, Muscatine, Iowa. In the nine years from June 30, 1936, to June 30, 1945, the rural consumers served in Iowa increased from 24,700 to 124,900. This electrical service is being provided primarily by privately owned utilities and by co-operatives, with a comparatively insignificant part by municipalities. Statistically, 39.8 per cent of the consumers are served by private utilities, 55.8 per cent by co-operatives, 3.2 per cent by municipalities, and 1.2 per cent by miscellaneous sources. Ten years ago an average monthly consumption of 100 kilowatt-hours per consumer was a goal toward which to work, whereas now the planning is in terms of from 300 to 700 kilowatt-hours per month per customer. An intense demand for electrical service by "practically every Iowa farm" is reported, including farmers who scornfully declined electrical service initially.

"Experience With Single-Pole Relaying and Reclosing on a Modern 132-Kv System" by J. J. Trainor (nonmember) and C. E. Parks (M '45) Public Service Company of Indiana, Inc., Indianapolis, Ind. In March 1941, the Public Service Company of Indiana placed in service, on an important 50-mile 132-kv intersystem tie line, a revolutionary installation of single-phase, or so-called single-pole, reclosing. This type of reclosing differs from the traditional three-pole reclosing in that, in the event of a single-phase-to-ground fault on the line, only the faulted phase is tripped open and then reclosed simultaneously at both terminals of the line. Experience has shown that for the short interval of time during which the faulted phase is open, either the parallel transmission line paths, or the ground itself, will assume the load current of the interrupted phase. Single-pole switching has permitted the transfer of large blocks of power over the lines so switched with less chance of system instability during faults which are confined to only one phase and ground, a condition which accounts for some 75 per cent of this company's transmission line faults for the type of line involved.

"Calculation of Power System Losses by Formula Using Constants Determined by the Network Analyzer" by J. B. Ward (A '40) Purdue University, Lafayette, Ind.

It is often necessary or desirable to determine the losses in a power system network for the purpose of forming operating schedules, for system planning, for interchange billing, or for other reasons. Ordinarily, the calculation of losses in a complicated power network is a lengthy and arduous task, perhaps involving a network-analyzer study of various operating conditions to determine the power flow and subsequently the making of line-by-line calculation of the I^2R losses. This paper presents a method whereby a formula for a given system is derived from a special network analyzer test, and the formula subsequently used to determine directly the total transmission loss for a large variety of operating conditions without further use of the analyzer. In this loss formula, the principle of superposition is utilized to let one set of analyzer studies serve to establish the line flow for a wide range of generating schedules with a given set of loads. The formula expresses the total loss as a function of station and tie-line kilowatt and kilovar loadings and corresponding bus voltages, information which is obtainable readily from station log sheets. The accuracy of this method is difficult to evaluate in general terms, inasmuch as errors depend to a large extent upon the assumptions and simplifications made in any specific case.

"Trends in Rural-Line Sectionalizing" by R. F. Quinn, (M '44) of the General Electric Company, Schenectady, N. Y. The enormous growth of farm use of electricity carries with it not only the responsibility but the economic necessity of assuring to farms a high quality of electric service. An electrified farm is an industry and it must be able to rely upon dependable electrical service. Ideal protection of rural lines would be obtained by the proper combination of oil circuit reclosers and fuse cutouts. The first step would be the application of enough oil circuit reclosers to handle all nonpersistent faults occurring anywhere on the system. Because the protective orbit of a recloser normally covers an extensive part of the system, this application of reclosers would result in relatively long sections of main lines, and would leave most branch-line taps unsectionalized. The second step, therefore, would be the application of enough fuse cutouts to supplement the reclosers by dividing the system into relatively short sections and by protecting all branch taps, so that a persistent fault would be isolated and its effects limited to a decidedly smaller section of the system that would be possible by reclosers alone. Proper co-ordination between the reclosers and the fuses would be required to assure that: (1) the reclosers handle all nonpersistent faults anywhere on the feeder; (2) the fuses do not blow on nonpersistent faults; (3) the fuses do blow to isolate persistent faults. Dependable co-ordination of this kind has been obtained on urban feeders, using station-type circuit breakers controlled by induction type vary-inverse-time-overcurrent relays. What the industry needs today is a recloser for rural line applications that will co-ordinate with fuses with the reliability of performance of station-type equipment.

CONFERENCE ON INDUSTRIAL POWER APPLICATIONS

John M. Webb (A '35) head of the department of electrical engineering, Eli Lilly and Company, Indianapolis, Ind., presided over the conference on industrial power applications. In general it could be concluded that electronic control of motor drives in conjunction with individual machine drives, is going to be more and more important to the industrial electrical engineer. The protection of plant equipment and distribution systems from lightning, and means of providing a flexible power distribution system also will receive considerable attention.

"Multiple Generator for Paper Machines" by G. E. Plaisted (A '46) General Electric Company, Schenectady, N. Y. It definitely is recognized that paper and paper products play a large part in our economics today. The usage is a bench mark to determine the level of our standard of living. The paper and paper products industry provides almost 400 pounds of this material annually per capita in the United States. This is several times the amount available in any other country. The potentialities of paper and paper products began to be recognized shortly after World War I. To increase production required more machines or machines of higher speeds or greater width. Chemical engineers developed processes for preparing the cellulose fibers in greater quantities and from materials heretofore considered inferior, and power engineers were faced with the problem of providing means of driving wider machines at higher speeds. For economic reasons, a paper machine was required to operate over wide speed ranges in order to make various weights and grades of paper. Mechanical drives involved line shafts, combinations of cone pulleys and belts for speed adjustment of individual paper machine sections, and gears and clutches for obtaining exact roll speeds and providing means of starting and stopping individual sections. Electrical engineers recognized the possibility of driving individual sections by individual motors, thereby eliminating heavy mechanical parts. Because its speed is a function of field flux and applied voltage, the d-c motor had inherent characteristics which made it particularly adaptable to a co-ordinated drive. The designers problem became one of maintaining an accurate speed relation among each of the several motors involved. The early sectional drives utilized one generator with individual motors, their speeds being maintained by field control. A further improvement in performance can be obtained by using an individual generator for each motor.

"Lightning Protection for Industrial Plants" by A. M. Opsahl (M '41) Westinghouse Electric Corporation, East Pittsburgh, Pa. On the basis of statistical averages, an industrial structure from 60 to 100 feet high and 100 by 100 feet in ground plan standing above a level plane may expect to receive a lightning discharge once in four or five years, or, in other words, one in some five such structures may expect a discharge each year. Surrounding objects will reduce the probability of direct hits, and the structural damage will depend upon the nature of the structure and protective installation if any. The installation of electric equipment used in industrial service usually is of the dry type with low factors of strength over normal voltage ratings. This fact makes it especially desirable with reference to industrial installations to use lightning rods and diverters to carry off direct strokes and surges, and to use lightning arresters specially designed and adapted to industrial service.

"Automatic Contouring Control" by J. M. Morgan, Jr. (nonmember) General Electric Company, Schenectady, N. Y. When used with the proper type of machine tool, this special type of control makes possible the machine duplication of work pieces of odd shapes by duplicating or reproducing the contour or shape of a master pattern or template. The electronic contouring control system described at Indianapolis combines basic electronic and magnetic circuits into a wholly electric and highly accurate system which provides constant speed of tool travel, great flexibility of control of machines, and continuous instead of step-by-step control. The electronic contouring control system operates as a "closed loop" system or, more specifically, as a "positioning follow-up" control system. Signal voltages generated in the tracing head by deflection of the tracing head stylus are used to operate the system. These signal voltages are fed into the contouring control desk where they are amplified and operated upon. The two resulting direct output voltages are fed into the motor-control circuits which in turn control the corresponding feed motors, which drive the tracing head—as well as the tool—so that the stylus follows the edge of the template. As there is a fixed mechanical relationship between the tool and the tracing head, and also between the work-piece and the template, the tool will duplicate on the work piece the contour of the template.

"A Demonstration of Industrial Distribution Systems" was given by H. G. Barnett (M '41) Westinghouse Electric Corporation, East Pittsburgh, Pa., displaying the comparative characteristics of radio and network distribution systems for industrial plants. The demonstration showed graphically, through the operation of a simulated industrial-plant system arranged on a panel board, that although the radial type of distribution system is less expensive to install and somewhat simpler physically, the network type of plant distribution system introduces an element of flexibility and freedom from interruption shutdown that very well may more than justify the initial instrument of expense. With the network system, any faults occurring in plant units are isolated quickly without interrupting other units; faults occurring between plant units are isolated without interrupting any units; repair, maintenance, and testing of elements of the plant distribution system can be accomplished at any time—even a transformer can be replaced—without interrupting the operation schedule of the plant.

CONFERENCE ON COMMUNICATIONS

The communication session was devoted to radio communication except for one paper on networks containing vacuum tubes. J. J. Pilliod (F '34) American Telephone and Telegraph Company, New York, N. Y. was chairman of the communications section.

"Manufacturing Problems of Air-borne Transmitter Equipment" by J. A. Green (A '42) Collins Radio Company, Cedar Rapids, Iowa. The various problems stemming from the peculiarities of aircraft installation lead to certain manufacturing problems. Factors of vibration, altitude which affects temperature and pressure, and power supply all must be taken into consideration.

"Phenomenon of Microwaves," a visual demonstration by H. R. Gruelle (nonmember) Indiana Bell Telephone Company, Indianapolis, Ind., supplemented the paper on the microwave relay system by showing some of the characteristics peculiar to microwaves.

"Analysis of Four-Terminal Networks Containing Vacuum Tubes" by Wilton R. Abbott (A '40) Iowa State College, Ames. Analysis and synthesis of four-terminal networks and systems of networks when these networks are linear, passive, and bilateral has been greatly facilitated by their study as a class without regard for their actual internal structure. It would seem that such a treatment of networks containing vacuum tubes would be of equal value. The method consists of determining four independent parameters of the circuit. Choice of equations depends on circumstances. In some cases if the complete circuit can be broken up into several interconnected four-terminal networks, it undoubtedly will be easier to determine the characteristics of each element and combine them.

"Mobile Radiotelephone Service" by J. G. Harden (nonmember) Indiana Bell Telephone Company, Indianapolis, Ind. Many of the requirements involved in providing the radio portions of mobile radiotelephone systems for use in common carrier service are similar to those encountered in city and state police and power company systems. Additional problems are imposed on a system which must provide satisfactory communication between mobile units and instruments in a regular wire network. Most of the special requirements of such a system are incorporated in a control terminal which provides for combination of two radio-derived one-way transmission paths into a single two-way circuit; combination of various receiver outputs delivered to the terminal by wire lines; introduction of fixed gain as required by transmitting and receiving branches; introduction of automatic gain in transmitting branches to maintain optimum modulation; provision of necessary features for switchboard operation; monitoring of emissions and frequency of the stationary transmitter; and provision of various test functions. These requirements are those of a primarily urban system.

"A Multichannel Microwave Radio Relay System" by H. S. Black (F '41), J. W. Beyer (M '36), T. J. Grieser (nonmember), F. A. Polkinghorn (M '39) Bell Telephone Laboratories, Inc., New York, N. Y. An eight-channel microwave relay system used by the Army and Navy uses frequencies approaching 5,000 megacycles. The waves are concentrated into a sharp beam and do not travel along the earth much beyond seeing distances. Because the waves are sharply focused other systems in the vicinity can operate on the same frequency and only one four-millionth of the power required by a nondirectional antenna is needed. Although this equipment was designed for military use its basic principles and design features can be applied to telephone and telegraph communication. Sharply beamed radiation and reception, pulse modulation, and time division multiplex are all factors which make the system attractive in competition with other communication systems. *Electrical Engineering* November 1946, volume 65, page 516, carries an article on this subject by F. B. Bramhall (F '44) The Western Union Telegraph Company, New York, N. Y. on "Radio Relays for Telegraphy."

CONFERENCE ON INDUSTRIAL CONTROL

Two papers which evoked especial interest at this meeting were the ones on induction motor control and a-c magnets. The paper on induction motor control was apparently based on much development work and represents one approach to the problem of variable speed a-c motor drives. Possibilities which drives of that general nature offer undoubtedly will lead to further development involving induction motors under unbalanced conditions. The paper on a-c magnets presented much information with regard to a-c magnet performance which is not generally known.

Session Chairman R. W. Jones (A '42) associate professor of electrical engineering, Northwestern University, Evanston, Ill, stated that the gap which exists between textbooks and periodical literature is a difficult hurdle for the young engineer and that it might be advisable for the Institute to sponsor sessions of an educational nature on a variety of specialized topics at the annual meetings. He also observed that there is little material available on a-c magnet design outside the field and that the Institute could well serve the members if publication of such material could be arranged.

"Trends in Co-ordinated Control" by G. A. Moffett (M '44) General Electric Company, Schenectady, N. Y. Ranking very highly in the long list of control activities are "co-ordinated control equipments" which have offered great improvements through increased output in continuous processes such as for paper, textiles, rubber, and steel. With mechanical systems comprising single motors, line shafts, gear reducers, couplings, and power take-offs, the devices for adjusting speed, entry, wind-up, tension, and so forth are not self-regulating and frequently must be readjusted manually. The electronic sectionalized drive has replaced the mechanically coupled system and makes possible a more simplified and flexible setup which is self-regulating and continuously adjustable. Co-ordinated control equipment has many applications other than sectionalized drive. These include large wind tunnels with thousands of horsepower on a single shaft, model testing, power absorption and recovery systems, chemical and petroleum processes, printing presses, and automatic furnaces. During the war a large number of control systems were devised to accomplish certain functions. The experience gained has enabled the best systems to be chosen for specific applications. The next move is one of standardization of functions and their circuit components without hindrance to flexibility of application.

"Control of Slip Ring Motors by Means of an Unbalanced Primary Voltage" by N. L. Schmitz (A '43) Cutler-Hammer, Inc., Milwaukee, Wis. Present applications of the unbalanced primary speed control for slip ring induction motors include crane and drawbench installations where a high lowering or return speed and braking action for slowdown is desired. To obtain the required unbalanced voltages for operation an autotransformer suitably tapped is used. As is true of any apparatus, this equipment has limitations. Most talked of limitation is heating of the motor. Current values of 200 to 250 per cent are encountered in the primary phases at slow speeds, but this is also true of armature loop currents in d-c motors. Even with a more severe duty cycle than the 15 seconds on and 45 seconds off specified by the American Standards Association undue heating is improbable, because coils adjacent to the hot winding are cooler thus facilitating heat transfer, and temperature equalizes rapidly when the motor is operating with balanced currents under load or when it is at rest. This is claimed to have proved true in tests as well as in theory. A wider speed range is possible if capacitors are used in the secondary circuit in place of the usual resistors. Use of inductance in series with the capacitance in the secondary permits further reduction in speed. In order to obtain large secondary current at high power factor for maximum torque at low speed the values of inductance and capacitance are selected to approach series resonance at the frequency of induced secondary currents.

"Problems in Design of A-C Magnets and Solenoids" by L. T. Rader (M '43) Illinois Institute of Technology, Chicago. Although a-c magnets have existed for some time, little has been published about their design. An attempt is made to present for consideration some of the difficult problems involved. To facilitate a strictly analytical solution for the design of a-c magnets several assumptions are usually made: flux varies directly with current, saturation is disregarded; leakage flux is neglected; iron losses are neglected; voltage drop is neglected. These assumptions are so bad that a solution dependent on them is very misleading. In order to design a magnet certain specifications must be furnished the designer. He must know the load displacement curve, duty cycle, quiet pull required in the sealed position, mechanical life, dimensions to which the device must conform, and the maximum volt-amperes which can be drawn from the line. The design problem consists of two parts—conditions when the magnet is closed or sealed

and conditions when the magnet is open. The problems encountered with the magnet in the closed position are all involved with those factors producing heat. Consequently voltage and steel are important in these considerations. For the open position force of the magnet is a major consideration and the leakage path assumes great importance. Few data are available on factors affecting leakage and volt-ampere data are used as a substitute. A rigid requirement of magnets in the closed position is that they must be quiet. Saturation is involved in this problem and it immediately deviates from a rigid analytical solution. Also important is the mechanical life. Rapid acceleration which results from the device being able to operate at 85 per cent voltage produces high impact forces. Repeated application of such forces introduces many problems of mechanical nature to design.

CONFERENCE ON BASIC SCIENCES

A variety of subjects was covered in the session on basic sciences, of which K. W. Miller (M '29), Armour Research Foundation, Chicago, Ill., was chairman.

"The Electronic Numerical Integrator and Computer" by T. K. Sharpless (A '44) University of Pennsylvania, Philadelphia, Pa. Need of the United States Army Ordnance Department Ballistic Research Laboratories at Aberdeen, Md., for a machine capable of rapidly performing calculations for preparation of firing tables initiated work on the electronic numerical integrator and computer. The machine, built at the University of Pennsylvania, Moore School of Electrical Engineering, was made sufficiently flexible to enable it to be applied to a variety of problems other than exterior ballistics, which involve numerical integration. Considerable work is required to prepare problems for solution but once the mathematicians have done this the machine performs the computations extremely rapidly. Comprising the machine are 20 accumulators, which perform the functions of addition, subtraction, and storage of numbers; 3 function tables, which permit looking up of functional values of empirical data; a high-speed multiplier; a divider-square rooter; a constant transmitter, which works with an International Business Machines card reader; and a transmitter which uses an IBM card punch. Also provided are the master programer, cycling unit, and initiating unit which control the operations of the machine.

"Radioactive Isotopes and Their Applications" by G. Freidlander (nonmember) General Electric Company, Schenectady, N. Y. Radioactive tracers have been employed for several purposes which include study of the behavior of minute quantities of material; study of exchange reactions; tracing of an element through chemical changes, which is useful in organic syntheses, analytical separation processes, and a multitude of biological researches; and for following migration of material. The latter purpose has found use in diffusion studies, investigations of corrosion and wear, and applications to process control. Limitations in application are determined by the half-life of the isotope, the chemical form and amount in which it is available, the type and energy of radiation, and the activity per unit weight necessary. The method by which an isotope is produced must be considered because it usually determines the available specific activity.

"Latest Developments in High Energy Physics" by Karl Lark-Horovitz (nonmember) Purdue University, Lafayette, Ind. As an outcome of war developments many physicists have become interested in electronic devices, and particularly a great many physicists whose main interest was and is in the field of nuclear and high energy physics or cosmic rays. These men, using techniques developed during the war, have shown that new devices can be built using such engineering methods as have been available in the high-frequency fields in radar development. Applying these ideas successfully to the acceleration of particles, they designed and built devices which promise to produce high energy protons of the order of 40 to 100 million volts in the so-called linear accelerator, high energy electrons in a device which is called the synchrotron, and high energy deuterons in the frequency-modulated cyclotron. These developments promise to open an entirely new field in nuclear physics research.

"Solar Electron Radiation and Its Effect on Power Transmission and Communications" by Doctor J. T. Wilson, Allis-Chalmers Manufacturing Company, Milwaukee Wis. Almost everyone has witnessed the effects of sunspot activity in disruption of radiobroadcasting during magnetic storms. The effect of solar electron radiation has had its effect on power transmission as well.

National Electronics Conference Papers Digested

For the purpose of giving AIEE members a brief cross section indicating the scope and nature of the 57 scheduled technical subjects discussed at the second National Electronics Conference in Chicago, October 3–9, 1946, the following brief digests are presented, drawn primarily from information furnished by the authors. Except to the extent that individual authors may have extra copies of their papers these papers are not individually available. All these papers are scheduled to be published in the "Proceedings" of the second National Electronics Conference, which currently is in production and is expected to be available for distribution early in 1947. Persons interested in securing copies of the "Proceedings" should take action immediately, forwarding a request *and* a remittance of $3.50 to E. H. Schulz, secretary, National Electronics Conference, Inc., Technology Center, Chicago 16, Ill.

TELEVISION

"Color Television—Latest State of the Art," by P. C. Goldmark (M '45) Columbia Broadcasting System, was presented as an oral dissertation covering calculations and performance characteristics of the equipment and procedures currently involved in developing color television.

"Westinghouse Color Television Studio Equipment" by D. L. Balthis of the Westinghouse Electric Corporation, described equipment patterned after the color television studio equipment designed and developed by the Columbia Broadcasting System in New York, N. Y. The studio equipment consists of the electrical and optical equipment required to convert a 35 millimeter color slide, or a 16 millimeter color film and its associated sound, into signals suitable for input to an ultrahigh-frequency color television transmitter utilizing three primary colors. Sound and picture signals are transmitted on the same carrier frequency, with special arrangements for color synchronization.

"Television Transmitter for Black-and-White and Color Television" by Norman Young of the Federal Telecommunication Laboratories was descriptive of the equipment indicated.

"Stratovision System of Communication" by C. E. Nobles of the Westinghouse Electric Corporation and W. K. Ebel of The Glenn L. Martin Company was descriptive of the experiments and plans that are being directed toward the development of a series of air-borne television transmitting stations intended to serve most of the United States from a small number of points. It is contemplated that a network of eight airplanes flying at a height of 30,000 feet and some 400 miles apart will be able to perform a service that would require 100 ground relay stations 35 miles apart.

"The Electrostatic Image Dissector" by H. Salinger of the Farnsworth Television and Radio Corporation describes a tube which uses magnetic focusing but which achieves scanning by electrostatic rather than magnetic deflection. The theory of this tube as worked out indicates that distortions are greatly reduced as compared with the conventional method of operating a dissector. The scanning generators are of the so-called "bootstrap" type. Deflection is accomplished by means of several wires (12 in the present tube) disposed axially inside the tube. The two scanning sawtooth voltages are applied in different proportion to each of these wires.

"The Use of Powdered Iron in Television Deflecting Circuits" by A. W. Friend (M '39) of the Radio Corporation of America. The horizontal deflection of electron beams in television systems heretofore has required excessive energy dissipation and expensive circuit components. For deflection by magnetic means, the deflection transformer and yoke have presented serious problems in the economical design of television receivers. Now, low-loss systems have been constructed which require no additional electric energy and which provide large increases in deflection capability. Also, the development of transformer and yoke cores molded from powdered iron have resulted in appreciable cost reductions as compared with sheet or strip-metal types. Molded core structures produce negligible acoustic radiation in comparison with laminated core structures. A low-cost system has been constructed to provide deflection of a 27-kv electron beam, yielding maximum picture size on a 60-degree kinescope driven by a single 6BG6G beam tetrode.

"Television Equipment for Guided Missiles" by C. J. Marshall, Wright Field, Ohio, and Leonhard Katz of the Raytheon Manufacturing Company, outlined experiments with lightweight transmitting equipment begun as early as 1941 by the Army Air Forces at Wright Field.

ELECTRONIC INSTRUMENTATION

"A Method for Changing the Frequency of a Complex Wave" by E. L. Kent of C. G. Conn, Ltd. This method is applicable in tone analysis, tone synthesis, psychological studies in tone quality, and other similar studies. The frequency of a complex wave may be changed by the use of a special cathode-ray tube or by the use or ordinary radio tubes utilized in special circuit arrangements.

"Detectors for Buried Metallic Bodies" by L. F. Curtis (F '29) of the Hazeltine Electronics Corporation, briefly outlines some of the problems encountered in the development of a detector for the Engineer Corps of the United States Army under the supervision of the National Defense Research Council.

"The Pressuregraph" by A. Crossley of the Electro Products Laboratories, Inc. This device is an electronic instrument for indicating in a linear manner the static or dynamic pressure of an internal combustion engine, pump, or other pressure device. The Pressuregraph consists of a diaphragm upon which the pressure waves impinge, the diaphragm being part of an electric capacitor. Movement of the diaphragm results in change in capacitance which in turn actuates a bridge or electronics circuit causing modulation of a high-frequency carrier, the modulated wave being amplified and delivered to a cathode-ray oscillograph which in turn presents a visual translation of the pressure-time picture that is the indicator card of the device being investigated.

"The Notch Wattmeter for Low-Level Power Measurement of Microwave Pulses" by D. F. Bowman of the Hazeltine Electronics Corporation presents a review of the basic methods of microwave power measurement. In the "notch wattmeter," a method is employed whereby the radio-frequency pulse of unknown amplitude is matched in amplitude with an interrupted continuous-wave signal. The interrupted portion, or notch, of the continuous-wave signal is adjusted to equal in length and to coincide in time with the unknown pulse. This adjustment facilitates accurate matching of amplitudes. The amplitude of the combined signal, as measured by an average power-indicating instrument, therefore is an accurate measurement of the peak pulse power of the unknown radio-frequency signal.

"The Mechanical Transients Analyzer," by G. D. McCann (M '44) of the Westinghouse Electric Corporation presented a brief description of the Westinghouse Electric analog computer which was developed for the solution of complex algebraic and and differential equations. The differential equations specifying the physical system are represented by electric circuit parameters such as resistors, inductors, and capacitors. The variables of the system are represented by currents and voltages. Electronic amplifier circuits are used to represent negative impedance elements or energy sources, such as required for servomechanisms. Special electronic circuits and techniques have been developed to produce arbitrary steps of excitation functions and initial or boundary conditions.

"High-Performance Demodulators for Servomechanisms" by K. E. Schreiner of the Servomechanism Laboratory, Massachusetts Institute of Technology. As the frequency-response range of servomechanisms is extended, the necessity for minimizing time delays in the response of individual components correspondingly is increased. In many instances, in servomechanisms using a-c data transmission, one of the important time delays is that introduced by the demodulator—or phase-sensitive detector—which in addition to small time delay should have the additional characteristic of low ripple component of output voltage. Normally, either of these characteristics may be attained at the expense of the other. A particular high-performance circuit is described and analyzed, the principal feature of which is the combination of small time delay and low ripple components of output voltage. The detecting element is a double-triode, the grids of which are excited in such a way as to establish a bidirectional conduction path between input and output for a very short period of time while the signal voltage is passing through its peak. The unique feature of this type of demodulator is essentially that the output voltage is not allowed to change except at the instant of peak signal input.

"The Theory and Design of Several Types of Wave Selectors" by N. I. Korman of the Radio Corporation of America. A wave selector is a device for sampling, measuring, or injecting a traveling wave in a transmission line or wave guide. As a knowledge of the traveling waves on a transmission line is equivalent to a knowledge of the standing waves, and as the wave selector can measure the traveling waves without the use of any mechanically moving parts, it can supplement and in many cases replace the more complicated standing wave detector. The principles of operation of the wave detectors known as the long-slot, electromagnetic, reversed-coupling, and phase-indicating types are given, and applications discussed.

"An Oscillographic Method of Presenting Impedances on the Reflection Coefficient Plane" was described in some detail by A. L. Samuel (M '34) of the University of Illinois.

"Electron Optics of Deflection Fields" were described by R. G. E. Hutter of Sylvania Electric Products, Inc. The properties of electric and magnetic deflection fields commonly used in cathode-ray and television tubes are discussed and mathematical theory expressions for the field quantities are given. If the assumption be made that the deflection of electrons from the optical axis is small while the electrons are under the influence of the deflecting field, the problem of finding general mathematical expressions for the deflection and for the deflection defects can be solved by two methods—the "path method" and the "Ikonal-method"—which are known from the theory of electron lenses. The resulting general expressions may be applied to a number of special fields.

Trends in cathode-ray oscilloscope design were discussed by W. L. Gaines of the Bell Telephone Laboratories, Inc. War demands for the development and servicing of radar and other equipment, and future demands now foreseeable, require signal channels which include attenuators, delay networks, and amplifiers capable of handling transient wave forms, and precision slave start–stop sweeps with a high ratio of operate–restore time. Accessory circuits such as lockout, sweep delay, and timing circuits have been found to be extremely convenient for many purposes and essential for some purposes. Inherent difficulties include adequate delay networks and signal attenuators for the signal channel, and adequate unblanking for the sweep circuit. Also, cross talk between circuits within the oscilloscope requires careful attention to grounding paths and to by-passing as well as shielding.

DIELECTRIC AND INDUCTION HEATING

"Microwaves and Their Possible Use in High-Frequency Heating" by T. P. Kinn

(M '44) and J. Marcum discusses the sources of power now available in the microwave region (1,500 to 30,000 megacycles) giving available power output and other pertinent characteristics. Coaxial transmission lines and wave guides were explained in simple terms.

"Ignitron Converters for Induction Heating" by R. J. Ballard and J. L. Boyer (A '43) of the Westinghouse Electric Corporation described a new type of ignitron frequency converter for melting-furnace and forging-heater applications. Derived from the d-c to a-c parallel inverter circuit, which has been known and used for many years, this new circuit is called the "cyclo-inverter." Three-phase 60-cycle power is converted to single-phase power at a higher frequency by means of a single conversion.

"Dielectric Preheating in the Plastics Industry" by D. E. Watts, G. F. Leland (M '45), and T. N. Willcox (A '38) of the General Electric Company. Dielectric preheating is a comparatively new industrial tool in the plastic industry. By this means, heat is generated within plastics materials just prior to molding, by applying a high-frequency electric field. The "preform," thus being uniformly soft, allows new techniques to be used in the processing of thermosetting plastic materials.

"The Problem of Constant Frequency in Industrial High-Frequency Generators" by E. Mittelmann of the Illinois Tool Works is related to the recent frequency allocation for medical and industrial generators by the Federal Communications Commission which emphasized the problem of constant-frequency generators. The requirement for constant frequency poses two problems:

1. To maintain a constant frequency operation.
2. To force the generator to this frequency.

A solution has been found in the combination of a sharp-cutoff high-pass filter and a sharp-cutoff low-pass filter providing control voltage outputs of different polarities for frequency deviations above and below the assigned frequency.

INDUSTRIAL APPLICATIONS

"Large Electronic D-C Motor Drives" by M. M. Morack (M '42) of the General Electric Company describes the application of electronic control equipment to d-c motors of 600 horsepower. The electronic drive system provides control means including both the starting and the running of the drive motor. The functions are equivalent to the Ward Leonard system, in that an ignitron rectifier replaces armature voltage adjustments by field control.

"Electronic Speed Control of A-C Motors" by W. H. Elliott of Cutler-Hammer, Inc., reviews briefly some of the basic electronic circuits which may be employed to control the more common types of a-c motors, such as the drag-cup shaded-pole, universal, and wound-rotor motors.

"The Electronic Method of Contouring Control" by J. Morgan of the General Electric Company. Contouring control is basically the following of a master template or pattern by a stylus or tracer. Using signals generated by the tracer, other portions of the contouring system control the mechanism, so that the cutting tool reproduces the contour of the template on the work. This electronic system uses a new combination of basic electronic circuits to obtain the desired controls.

"Production Test Facilities for High Power Tubes" by W. L. Lyndon and B. Sheren of the Radio Corporation of America described an equipment and installation which was engineered and designed especially for the testing of power tubes. The physical size of high power electron tubes, plus the magnitude of voltages, currents, and power outputs encountered, necessitate the design of special equipment for the testing of these tubes. Flexibility in such equipment is essential.

ANTENNAS AND WAVE PROPAGATION

"Radio Propagation at Frequencies Above 30 Megacycles" by K. Bullington of the Bell Telephone Laboratories, Inc. Radio propagation is affected by many factors, including the frequency, distance, antenna height, curvature of the earth, atmospheric condition, and the presence of hills and buildings. The influence of these factors is represented in nomograms, by means of which an estimate of the received power and the received field intensity for a given point-to-point radio transmission path ordinarily can be obtained in one minute or less.

"Interference Between Very High Frequency Radio Communication Circuits" by W. R. Young, Jr., of the Bell Telephone Laboratories, Inc. Interference between different radio circuits is an old problem, one which generally has been solved in the past by trial and error and by "hand-tailored" special filters and other equipment. With the general increase in the usage of radio communication, the amount of potential interference is greatly increased. Generally there is a large difference between transmitting and receiving energy levels. As a result, spurious radiations, spurious responses, and lack of sufficient receiver selectivity may in many instances cause interferences. Common causes of such interferences are discussed. Sample measurements are given to illustrate the relative magnitude of the various modes of behavior. Formulas are given which permit computation of the frequency of these disturbances. A method is described for making charts suitable for a given type of equipment from which the spurious frequencies can be read directly as a function of the operating frequency.

"Aircraft Antenna Pattern Measuring System" by Otto Schmitt, Airborne Instrument Laboratories. This is a "packaged system" for measuring the radiation patterns of aircraft antenna by the use of scale models.

"Improvements in 75-Megacycle Aircraft Marker Systems" by B. Montgomery of the Northwest Airlines describes improved aircraft marker systems which permit air line pilots to follow their courses to their destinations with greater precision. The system consists of transmitters operating on 75 megacycles with antenna systems of a type which concentrate the radiation in an upward direction, the markers thus indicating the airplane position in relation to range stations on the ground.

"Problems in Wide-Band Antenna Design" were discussed by A. G. Kandoian (A '40) of the Federal Telecommunication Laboratories. With the increased use of higher and higher frequencies, the problem of wide frequency range operation of antennas of various types assumes more importance, and its solution becomes more feasible. The most general requirement is that both the impedance and radiation patterns be essentially independent of frequency over the operating range, whether applying to nondirective antennas or to highly directive antennas.

"Slot Radiators" were described and discussed by A. Alford (M '42) consulting engineer, Chicago, Ill.

"Results of Field Tests on Ultrahigh-Frequency (490 Megacycles) Color Television Transmission in the New York Metropolitan Area" were reported by W. B. Lodge of the Columbia Broadcasting System. Since January 1946, the CBS has operated an ultrahigh-frequency color television transmitter located on top of the Chrysler Building in New York City. Surveys utilizing mobile recording field intensity equipment and involving operation of actual television receivers at a large number of different locations throughout the metropolitan area, carried on continuously since the transmitter began operation, have provided considerable qualitative and quantitative data on:

1. The extent of the service area.
2. The effect of changes in height of a receiving antenna.
3. Shadowing.
4. Effect of trees and foliage.
5. Stability of transmission.
6. Location and extent of areas receiving multipath transmission.
7. Efficacy of methods of eliminating "ghosts" in areas receiving multipath signals.

A method of predicting possible areas of multipath transmission has been developed.

FREQUENCY MODULATION

This was a panel discussion on the design of frequency modulation receivers.

"A Permeability-Tuned 100-Megacycle Amplifier of Specialized Coil Design" was described by Z. Benin of Zenith Radio Corporation. Small coils were designed that could be positioned between tube and switch circuit components to make a compact radio-frequency tuner with a minimum of stray inductances. Powdered-iron-core permeability tuning improves the frequency drift with temperature. A variable-pitched winding technique was used to straighten out the tuning curve of the core-and-coil combination. The lengths of core-and-coil, the winding, and the powdered-iron materials were designed to give a logarithmic frequency travel curve. By mathematical analysis such a characteristic was shown to enable the elimination of trimmer capacity circuit adjustment and the substitution of an initial set-

ting of the tuning core. The result was a space-saving and frequency-drift improvement.

"Very High Frequency Tuner Design" by G. Wallin and C. W. Dymond of the Galvin Manufacturing Company describes the design of a tuner for use in the 88–108-megacycle frequency modulation band. The circuit used is essentially an adaptation of a single-oscillator double superheterodyne. One section of a dual triode tube performs the function of oscillator and first converter. The other section is used as the second converter. Three very "Hi-Q" elements used for tuning the antenna, oscillator, and variable intermediate frequency, are provided by means of permeability tuned transmission lines. This adaptation of the transmission line permits the use of an effective 250 micromicrofarad oscillator tuning capacitance, with attendant reduction in warm-up drift. A high order of temperature stability also is achieved.

"Front-End Design of Frequency Modulation Receivers" by C. R. Miner of the General Electric Company describes the methods used by the General Electric Company in solving the practical problems involved in the design of the front-end of a frequency modulation receiver, involving the use of a new type of variable inductance tuner, together with a unique mechanical design.

"A Single-Stage Frequency Modulation Detector" by W. E. Bradley of Philco Radio and Television Corporation describes the equipment and circuit combinations developed by that company.

"Frequency Modulation of High-Frequency Power Oscillators" by W. R. Rambo of the Airborne Instrument Laboratories. Practical cases arise in which it is desirable to frequency-modulate power oscillators operating at high frequencies, and thereby to avoid the use of amplifying and frequency-multiplying stages. A common-grid reactance-tube circuit is described that has an inherent tendency to reduce incidental amplitude modulation, permits the use of triode reactance tubes in the very high frequency and the ultrahigh frequency ranges, and has a configuration that permits its incorporation in ultrahigh frequency common-grid coaxial-line oscillator circuits. Lighthouse tube oscillators have been modulated over a plus or minus one-half per cent band, at 1,000 megacycles by this means; much wider band widths are possible in the very high frequency range.

AIR NAVIGATION SYSTEMS

"Automatic Radio Flight Control" by F. L. Moseley, Collins Radio Company, and C. B. Watts of the Federal Telecommunication Laboratories discusses the general problem of automatic control of aircraft flights on radio-defined tracks as an aid to point-to-point navigation, traffic control, and final approach and landing. A brief outline is given of the various radio navigational facilities which are available to define suitable tracks for automatic aircraft guidance, and the paper describes various systems which have been tested extensively. Widespread adoption of automatic radio flight control systems is recommended as an indispensable aid to all-weather operation of aircraft.

"Naviglobe—Long-Range Air Navigation System" by P. R. Adams and R. I. Colin, Federal Telecommunication Laboratories describes a directional transmitting type of system developed and proposed by the Federal Telecommunication Laboraties for the particular requirement of long-distance transoceanic air navigation. By virtue of its particular antenna disposition, and by using all quantitative signal relationships and not merely the conventional equisignal indications, the coverage is continuous and omnidirectional. Line-of-position indications are automatic and visual, corresponding to the true bearing of the observer, without the need of special charts. The fundamental features are designed on the basis that a minimum ground-station range of 1,500 miles is required to cover the North Atlantic area and the other likely transoceanic and polar routes. The study indicates that extremely reliable propagation at such ranges is to be expected most feasibly and economically if low frequencies in the neighborhood of from 70 to 100 kilocycles are used.

"Teleran—Air Navigation and Traffic Control by Means of Television and Radar" was presented by D. H. Ewing and R. W. K. Smith of the Radio Corporation of America. Wartime developments of radar technique offer a new approach to the problem of improving air navigation and traffic control, two fields in which existing equipment is obsolescent. Utilizing the special qualities of different special equipment developed originally for military service, the new proposed Teleran system presents aircraft position information to ground observers and controllers on a series of plan-position indicators. One indicator is used for each altitude layer, and is superimposed on a map of the region covered by the ground radar. This information, together with desired information of other factors, including weather and traffic control, is transmitted by television to each aircraft in the region. Each co-operating aircraft is equipped with a transponder beacon which serves not only to reinforce the radar echo, but also to provide an altitude-dependent reply which allows the ground station operators to differentiate among aircraft according to altitude.

RADIO RELAY SYSTEMS

"A Microwave Relay Communication System" by G. G. Gerlach of the Radio Corporation of America reviews the experimental results obtained with a 4,000-megacycle radio relay system connecting New York and Philadelphia. This system employs a frequency-modulated subcarrier which in turn is used to frequency-modulate the final carrier. Demodulation to the subcarrier frequency is effected at relay stations. Microwave relay equipment resulting from this experimental work which will be installed by the Western Union Telegraph Company in a circuit connecting New York, Washington, and Pittsburgh is described.

"Pulse-Time Multiplex Broadcasting of the Ultrahigh Frequencies" by D. D. Grieg (A '39) and A. G. Kandoian of the Federal Telecommunication Laboratories describes the result of operating experiences with a system in experimental operation in New York City. The system combines a total of eight high fidelity programs operating on the frequency of 930 megacycles; the pulse-time video band width utilization is 2.8 megacycles; base pulse repetition is 24 kilopulses per second per channel, with a time modulation deviation of plus or minus one microsecond.

"The Cyclophon—a Multipurpose Beam Switching Tube" by J. J. Glauber, D. D. Grieg, and S. Moskowitz (A '41) of the Federal Telecommunication Laboratories concerns the theory of operation of beam switching tubes in general, a type of tube originally developed for pulse-time modulated communication systems, but which proved to have many other interesting applications. The tube described employs secondary emission, and the paper discusses the characteristics of these dynodes and their equivalent circuits. Construction and processing of the tube are described in detail, and the results of life and vibration tests also are given.

MICROWAVE GENERATORS

"Continuous-Wave Ultrahigh-Frequency Power at the 50-Kw Level" by W. G. Dow (M '32) and H. W. Welch of the University of Michigan describes the theory and use of electronic apparatus capable of generating continuous-wave radio-frequency power at the 50-kw level in the frequency range from 350 to 600 megacycles. This equipment was developed originally during the war to meet certain military requirements for the jamming of German radar equipment.

"Microwave Frequency Stability" by A. E. Harrison of the Sperry Gyroscope Company. A review of the history of previous expansions of the radio-frequency spectrum emphasizes the importance of frequency stability in any region. The microwave region is no exception. Several methods are available for providing the frequency stability that is essential throughout the radio frequency spectrum, including the microwave region. Direct crystal control of the microwave power can be obtained by using klystron frequency multipliers and klystron power amplifiers. Automatic frequency control systems also may be used. The reference frequency may be either a crystal-controlled oscillator or a precision cavity resonator. All these methods are compared, and their advantages and disadvantages discussed.

"An All-Metal Tunable Squirrel-Cage Magnetron" by F. H. Crawford of Williams College describes the "donutron" which is tuned by the relative axial displacement of alternate anode segments through the flexure of one wall of the cavity in which the anode structure is supported. The best model to date tunes over a 1.5:1 ratio with power flat to 3 decibels. A single value of voltage and magnetic field is adequate for the entire tuning range. Of the various modes of operation, two are important: A long-wave tunable cavity mode, and a short-wave resonant re-

entrance-line mode. The former can be suppressed entirely, and the latter can be enhanced by a special phase-reversing anode. In the line mode, the best powers have been observed. In cold tests, modes of 4 centimeters as well as indication of even shorter resonances have been found.

"Design of Wide-Range Coaxial Cavity Oscillators Using Reflex Klystron Tubes" by J. W. Kearney (A '44) of the Airborne Instrument Laboratories describes the techniques utilized in the design of these tubes. The tuning range of each oscillator is approximately two to one in frequency in the 1,000- to 11,000-megacycle ranges. A survey of the general requirements of such oscillators for use as local oscillators in superheterodyne receivers and as signal sources in test equipment is followed by a discussion of the suitability of reflex klystron tubes for those oscillators in the particular frequency range. The operation of reflex klystrons is described with regard to their use over large frequency ranges, and adaptation of present-day tubes to coaxial resonators is discussed. Design characteristics are outlined for optimum cavity dimensions, suitable contacts to the tubes, noncontact-short-circuiting tuning plungers, and output coupling devices.

THEORETICAL DEVELOPMENTS

"Bunching Conditions for Electron Beams With Space Charge" by L. Brillouin of the Cruft Laboratory, Harvard University. A rigorous computation of electron trajectories within plane structures may be made, with the inclusion of space charge effect, by use of the Llewellyn method of integration. These computations can be used for a discussion of conditions leading to intercrossing trajectories that is bunching. The discussion is carried out in this paper for a conventional plane diode, for a diode wherein electrons are injected with a given initial velocity, for a velocity-modulated beam, for a plane magnetron, and for a plane magnetron with velocity modulation.

"Generalized Boundary Condition in Electromagnetic Problems" by S. A. Schelkunoff (M '34) of the Bell Telephone Laboratories, Inc. The conception of an idealized presently conducting boundary long has been helpful in simplifying electromagnetic problems. The paper deals with a generalization of this concept and particularly with its application to the theory of the magnetron.

"Conformal Transformations in Orthogonal Reference Systems" by C. S. Roys (M '45) of the Illinois Institute of Technology. Some general equations for conformal transformations which correspond to the Cauchy-Riemann Equations for Cartesian co-ordinates. These then are applied to shielding and to electron tube problems involving recurrent structures, leading to an equivalent, physically realizable unit structure. This removes the necessity of obtaining an appropriate transformation by cutting and trying.

INFRARED AND MICROWAVE SYSTEMS

"Reflex Oscillators for Radar Systems" by J. O. McNalley and W. G. Shepherd of the Bell Telephone Laboratories discusses problems encountered in the design of a series of reflex oscillators for military application. As an essential element of a radar receiver, a beating oscillator is required to heterodyne the received signals to intermediate frequency. Military requirements dictated the necessity of compact simple devices for such an oscillator. The reflex oscillator, because of its single-cavity tuning control and vernier tuning provided by the electronics tuning, provided a very satisfactory solution to these problems.

"Modulation of Infrared Systems for Signaling Purposes" by W. S. Huxford, of Northwestern University. In general, the sources used in "light" beam communication systems are modulated in one of three ways. The first method—mechanical modulation—utilizes shutters or mechanically vibrated elements to secure variation in intensity of the radiated beam after the beam leaves its source. In the second or optical method of modulation, some form of optical shutter such as a Kerr cell is used. The third method consists of varying the radiated flux generated in the source by electrical modulation of the power supplied to the lamp. A detailed description is given of the mechanical and optical methods of modulation used in captured German equipment. As an example of electrically modulated sources, the characteristics of a new source called the "concentrated-arc" lamp are described.

"Photo Detectors for Ultraviolet, Visible, and Infrared Light" by R. J. Cashman of Northwestern University considers photoelectric cell developments of the past few years, including photoemissive, photovoltaic, and photoconductive types. Photoemissive cells with caseium alloy cathodes have great sensitivity in the ultraviolet and the visible regions of the spectrum. Other cells of this type with pure metal anodes have considerable application in the ultraviolet region. Present-day photovoltaic or self-generating cells of the selenium type, most commonly used for work in the visible spectrum, are characterized by high output and good stability. Two new photoconductive cells of thallous sulphide and lead sulphide, which recently have been released, exhibit marked infrared sensitivity and excellent stability.

"Military Applications of Infrared Viewers" by G. E. Brown of the Engineer Board, Fort Belvoir, Va., described several items of infrared equipment developed by the Engineer Board. These items included the "Metascope," which employs an infrared sensitive phosphor for detecting sources of infrared radiation, and an infrared telescope using the 1P25 image tube for viewing areas which are "illuminated" by infrared radiation.

SPECTROSCOPY AND MEDICAL APPLICATIONS

"The Use of Radioactive Materials in Clinical Diagnosis and Medical Therapy" by J. T. Wilson of the Allis-Chalmers Manufacturing Company, described the methods for the production of radioactive isotopes and reviewed methods and equipment currently in use for inducing artificial radioactivity by means of bombardment with neutrons, alpha particles, protons, and gamma radiation. He explained the close relationship existing between the growing knowledge of nuclear phenomena and medical research pertaining to both diagnosis and treatment of human ailments. The use of radiophosphorus and radioiodine in clinical diagnosis and biological tracer techniques was reviewed. By the use of known and predetermined dosages of radioactive matter, it is possible now to trace accurately and definitely the transition and assimilation of such materials and of the materials with which they are mixed by both animal and vegetable tissue. With the door to an accurate knowledge of atomic structure opened by the results of recent nuclear research, and with the opportunity thus presented for discovering and studying the similarities, differences, and interrelationships of the atomic structure of animal tissue on the one hand, and of medicinal substances on the other hand, it now seems to the observer that medical science at last has the opportunity and the facility to become an exact science.

"The Mass Spectrometer as an Industrial Tool" by A. O. Nier of the University of Minnesota. Ten years ago, the mass spectrometer was but a complex laboratory device understood and used by only a few specialists working in highly scientific fields. Today it is an important tool in any laboratory or plant where gas analyses are required or where vacuum problems are encountered. It has been used as a means of making continuous automatic gas analyses of the process gas used in the large diffusion plant for the separation of uranium 235. It has been used also as a tool for readily and rapidly locating vacuum leaks in plant equipment of many varieties. The mass spectrometer is a gas analyzer depending in operation upon electronic principles. The gas to be analyzed is ionized by electron impact, and the ions thus formed are separated by magnetic fields into beams according to the weights of the ions. From the relative intensities of the various ion currents, the composition of the gas may be deduced accurately.

"The Cathode-Ray Spectrograph" by R. Feldt and C. Berkley of the Dumont Laboratories. This instrument produces complete spectral distribution curves of the source under test, at the rate of 240 complete spectra per second. The spectrogram is presented visually on a cathode-ray tube. The instrument was developed initially to facilitate the study of microsecond current pulses on a cathode-ray tube fluorescent screen, but the color of any other varying light source may be studied similarly. Provision is made for the insertion of transmission samples, chemical samples, or filters for calibration. Continuous recording may be accomplished through the co-ordinated use of a camera attachment, and indications may be given in positions remote from the analyzer when desired. Possible applications include: instantaneous color matching, fluorescent lamp characteristics, spectrographic research, continuous process control, and transmission

of spectral characteristics from remote or inaccessible locations.

RADIO AND RECORDING

"The Reduction of Background Noise in the Reproduction of Music From Records" by H. H. Scott (M '38) of the Technology Instrument Corporation. Conventional reproducing systems used with disk records or other noisy sources of audio-frequency signals have fixed transmission characteristics, regardless of the instantaneous requirements for reproduction of the applied signal with the optimum signal-to-noise ratio. By the use of "gate circuits" a system is achieved that is selective with respect to the type of applied signal, and which consequently tends to attenuate circuit types of signals while reproducing other types without substantial modification. Through proper design it is possible to obtain a system which will transmit with a high degree of fidelity those types of signals normally encountered in vocal and orchestral music, while at the same time discriminating against the usual types of background noises such as needle scratch and motor rumble. It also allows the reproduction of a greater frequency range than is common in conventional phonograph systems, thus reducing to a large degree the obvious difference between phonographic music and good frequency-modulation reception on a high grade radio phonograph combination. In its simplest form, the noise suppresser is reported as sufficiently inexpensive to be incorporated in a home-type phonograph, while the more elaborate version is suitable for use in broadcasting stations.

"Recent Developments in Magnetic Recording" by R. B. Vaile, Jr. (A '35) of the Armour Research Foundation. In the development of an ideal magnetic recorder some of the serious problems involved have to do with

1. The resolution of the recording head.
2. Distortion resulting from nonlinearity of the magnetization curve and other properties of the medium.
3. Constant speed drive.
4. Permanence of the record.
5. Transfer of the record from one part of the medium to an adjacent part.

The compromises which seem preferable in the present evaluation of solutions to these problems involve the use of

1. Longitudinal magnetization of the medium.
2. Speed between 0.5 and 5.0 feet per second.
3. Supersonic "bias."

In addition, extensive work has been done on the development of special materials, new test equipment, and special test methods.

MOBILE RADIO COMMUNICATION

This subject was discussed by a five-man panel, special attention being given to selective calling systems and to some of the specific items of equipment involved. "Signal Systems for Improving Railroad Safety" were discussed by K. W. Jarvis (M '34) consulting engineer of Winnetka, Ill.

NUCLEAR PHYSICS

"Some Fundamental Problems of Nuclear Power Plant Engineering" were considered by E. T. Neubauer of the Allis-Chalmers Manufacturing Company. As the temperature of an atomic reacting pile may range from a few degrees to millions of degrees, some of the important problems in the development of atomic power for commercial use pertain to the control of pile temperature and to the development of materials capable of withstanding ultra-high temperatures. Another important problem is that of adequate shielding for the protection of personnel. On the basis of best present knowledge, regardless of the size of the pile, the total weight of shielding must be of the order of some 30 to 50 tons or more to suppress dangerous radioactive emanations. Incidental to any commercial development, therefore, it seems probable that most pile controls will be remote so as to keep operating personnel at safe distances. Further, indirect heat transfer methods must be devised to permit heat to be transferred from the pile to utilization or conversion equipment without a corresponding transfer of dangerous and undesirable radioactive emanations. The per-unit cost of the energy available today (about 1/10 of one per cent of the theoretically possible total) from fissionable material currently available is within reason and comparable with that of ordinary combustible fuels. The cost and complexity of conversion and utilization, however, impose other problems. It is to be expected that the use of atomic energy for power purposes probably will find its first application in naval vessels and then in commercial ocean-going vessels. Some observers consider that this is a possibility, if not a probability, within the next five years or so. According to the author "it may be possible that the further divisions of the nucleus will reveal that the building blocks of the atom are the electron and the positron, instead of the electron, positron, and neutron; if so, there is a small possibility that some day science will learn the art of directing the nuclear parts in some special electronic tube or conductor. It is worth a gamble, for if this can be discovered, the greatest possible use and the highest possible efficiency for nuclear energy will have been found."

"An Accelerator Column for Two to Six Million Volts" was presented by R. R. Machlett. The direct acceleration of electrons and positive ions to energies of several million volts, using the electrostatic constant-potential generator, has become an important physical technique with applications to nuclear research, deep X-ray therapy, and the X radiography of heavy metal sections. Multisection vacuum tubes in which charged particles are progressively accelerated and focused, have been developed for such application. These accelerator columns consist of a multitude of planary electrodes spaced three per inch and separated by glass insulation. For a five-megavolt tube the column is 110 inches long and contains 330 electrodes.

"The Betatron Accelerator Applied to Nuclear Physics" by E. E. Charlton (A '45) and G. C. Baldwin of General Electric Company described applications of another particle-accelerating device. Splitting or fission of the uranium atom by X rays takes place most effectively at energies of 20 million electron volts, but at higher energies processes other than simple fission occur, and their nature currently is being studied with the aid of the 100-million-volt betatron. The nature of these other processes of higher energies is inferred from studies with elements lighter than uranium. For example, a nitrogen atom disintegrated into hydrogen, helium, lithium, and several neutrons. Warning that further significant advances in the science of the atomic nucleus will depend upon discoveries in fields not yet explored, Doctor Baldwin stated that "physicists today are hardly nearer an understanding of the atomic nucleus and of the nature of the particles which compose it than they were several years ago, even though the successful development of the atomic bomb may lead one to think that the fundamental questions of nuclear science all are answered. On the one hand, the radiation or absorption of energy by the electrons in a radio antenna, the flow of electrons in a vacuum tube, and the motion of electrons in an atom all can be described with high precision by well-established physical theories based upon simple laws of the electromagnetic fields of elementary charges of electricity. On the other hand, many nuclear phenomena are an almost unsolved riddle. The neutrons and protons which compose a nucleus are held together by extremely strong forces of very short range. We do not know the laws by which these forces act. We do not know what relationships exist, if any, between the nuclear and the electromagnetic fields. We know very little about the elementary particles of the nature of themselves—why there are so many kinds of them, how they interact with each other, whether they have internal structure, why some of them are radioactive, and so forth."

This 5-kw magnetron operates at 1,050 megacycles and is a General Electric Company product

INSTITUTE ACTIVITIES

Winter Meeting Papers

Presented in Four Main Groupings

An all-inclusive professional program and diversified social activities are promised for the AIEE winter meeting, January 27–31, 1947, in New York, N. Y. On the social side, as usual, a smoker and dinner dance will be held. Theater tickets will be available for out-of-town members, and special entertainment for women will be provided. Inspection trips will be arranged to places of interest and near-by industries. On the professional side technical sessions and conferences, segregated in four broad groups—industry, power, communication and science, and general applications—will cover practically the entire field of electrical engineering. The general session will be the common meeting ground of all four groups. During the meeting, presentation of the Edison, John Fritz, and Hoover Medals will be made. Details of the awards will be announced later. Meeting headquarters will be in the Engineering Societies' Building.

INDUSTRY GROUP

In the industrial field conferences will be held to highlight problems in the textile industry and to discuss the optimum industrial voltages for distribution systems. A session on industrial instrumentation is in prospect, and still another session will present a variety of industrial power applications.

For the electric welding people a conference will be arranged to discuss the application of resistance welders and the problems of power supply. A session on resistance welding will deal with battery welding, the secondary impedance of resistance welders, and calculations on the effect of introducing ferromagnetic welding materials into the throat of resistance welders, all of which are factors affecting the design of these machines.

Metallic rectifiers, which have come into extended use in many industries since the war, will be the subject of a conference. The proposed standards and the proposed test code for metallic rectifiers will be presented, together with other papers on this subject, and interested groups in the Army, Navy, and Air Corps will be invited to attend.

Industrial control devices will be the subject of another session, and papers will be presented on the design details of components, as well as on applications circuits of complete controllers. Also of interest to industrial engineers are a session in the communication and science group, dealing with the theoretical aspects of servomechanisms and a conference on electronic instruments.

POWER GROUP

A timely session is being held to revive interest in hydroelectric systems. A conference on interconnected systems is being arranged, as new problems have arisen in the five-year interlude since the last session on this subject. Considerable progress has been made in the solution of these problems, which makes the subject pertinent at this time.

In the field of power transmission and distribution, sessions are in prospect to present the latest information on lightning surges, grounding, and protective apparatus. In a session on electric machinery consideration will be given to the design of capacitor motors, the new NEMA standards for fractional horsepower motors, as well as the effect of steep front waves and chopped waves on transformers, and short-circuit standards for three-winding transformers.

COMMUNICATION AND SCIENCE

In the field of communication three sessions and a technical conference will be arranged. One of the sessions will deal with rural telephone facilities. The problem of providing telephone service to more people living in rural areas is now being attacked vigorously from many angles. Papers will be presented on each of the many approaches being used. Included will be papers on the design and application of carrier telephone systems for operation over rural power lines, the joint use of pole lines for power and telephone service, new methods of telephone wire construction for rural areas, and the use of radiotelephone to provide rural service.

Another session will be devoted to television. With more television stations going on the air and television receivers reaching the market in ever greater numbers, television is in the forefront of engineering advances and many aspects of its progress will be presented at this session. Included will be papers on transmitter and camera equipment, television receiver design, color television, and networks for television, both coaxial cable and microwave radio.

The session on communications components and techniques has scheduled papers on a variety of interesting subjects, for example, papers on recent improvements in telephone transmission systems, modern ocean cable technique, new developments in telephone switching systems, and subjects related to magnetic recording.

The technical conference in this group will be on testing telephone receivers and transmitters. Brief talks will be given by representatives of many concerns and institutions interested in this matter, and provision will be made for discussions and questions from the floor.

Applications in several fields of engineering will be emphasized at the session on instruments and measurements. Papers on new moving magnetic instrument design and magnetometer design for measurement of magnetic fields will stress wartime techniques as applied to peacetime uses. Other papers describe compensation for transformer loss in metering, measurement of crest voltages in ignition systems, and spectrographic measurements of light in rapidly changing spectra by use of a cathode-ray oscillograph for control or measurement purposes.

The conference on electronic instruments will consider a variety of applications, such as an electronic null detector for use with impedance bridges, pulse echo measurements on telephone and television facilities, defect location of materials by supersonic waves, and a general review of the electronics measuring field.

In the field of servomechanisms another session will attempt to answer some of the technical questions raised during the summer meeting in Detroit last June and to present more of the recent advances. A conference on servomechanisms also is proposed for the purpose of bringing together representatives of the technical personnel working in this field and representatives of the industrial people. An effort will be made to bridge the gap between the point of view of the technical man and the point of view of the nontechnical user of servomechanisms.

In the field of basic sciences a conference on applied mathematics has for its purpose the bringing together of mathematicians and engineers to discuss different engineering problems which have arisen during the past few years concurrently with new developments in the engineering and scientific fields. A session on topics of wide appeal in the basic science field will be held, and a conference is planned on the subject of the use of large-scale computing devices.

Many papers will be presented in various sessions on applications of electronics. In addition, a session on electronics will present a group of papers on rectifiers.

GENERAL APPLICATIONS

In the field of land transportation three sessions will be devoted to broad aspects of the subject. More specifically a session on light traction will deal mainly with the rapid transit system and subway equipment for New York, N. Y. Major General Charles P. Gross, chairman of the Board of Transportation, will give a special address at this session. Another session on electric railways will consider railroad electrification, power conversion and transmission costs, current collectors for high-speed service, and the modern Presidents' Confer-

ence Committee street car. Recent developments in various phases of Diesel-electric locomotives also will occupy a session.

Aircraft safety will be the subject of an all-day conference to discuss the factors affecting the safe operation of electric apparatus. Discussions will be prepared on assigned topics divided among the air line operators, the air frame manufacturers, and the accessory manufacturers. The program will be divided in six panels, as follows: safety, history, and requirements; types of electrical failures; protection in the electric system; application considerations; the safety program; and general discussion. The entire program will be conducted from a noncritical point of view with an attempt to bring out only constructive contributions.

The committee on safety will sponsor a symposium on grounding of portable equipment prepared by the committee on article 250, grounding, of the electrical committee of the National Fire Protection Association, for consideration in the development of the National Electrical Code; and discussions of the report will be presented by producers and users of portable electric equipment and by producers, installers, and users of the connecting cord and plug equipment involved in such grounding.

The enormous increase in the use of portable electric tools and other equipment in manufacturing establishments makes it highly important that engineers concerned with the development, use, and maintenance of such equipment be well informed as to the need for safeguarding and methods of accomplishing it.

Quality control will be discussed in a conference jointly sponsored by the American Society for Quality Control and the Institute's subcommittee on applications of statistical methods. The general theme of the conference will be applications to consumer goods. G. D. Edwards (M '26) president of the American Society for Quality Control, will speak on the purposes and progress of the society. This will be followed by two papers dealing with special applications of quality control to consumer goods. The conference will close with a general discussion, including all aspects, with particular reference to certification of consumer goods.

Further details of the entire program will be announced in the January issue of *Electrical Engineering.*

New AIEE Publication Policy Becomes Effective in January 1947

By Charles F. Wagner, Chairman AIEE Publication Committee

One of the fundamental purposes of the Institute is the dissemination to its members of information concerning new theories, new apparatus, and new developments throughout the electrical field. Incidental to its responsibility under this objective, and pursuant to action of the board of directors (*EE, June 1946, p 270*) the AIEE publication committee has been developing ways and means of improving the Institute's publications policy and procedure with the objectives of better serving the greatly enlarged and rapidly growing membership and of covering adequately the active and ever-growing electrical field.

The purpose of this report is to describe and explain the modifications in present publication policy and procedure constituting the new AIEE publication policy, which is to become effective immediately with 1947 publications. Before discussing the new policy, it is desirable to review the present policy which now falls somewhat short of membership needs, and to define the various kinds of publication materials that come to the Institute for consideration.

PRESENT PUBLICATION PLAN

The present AIEE publication policy was established in 1940. Most of the material published by the Institute is submitted as technical papers. These papers are referred first to the technical program committee for review and approval; if approved, they then are scheduled for presentation at a national or District meeting, where they are open for discussion.

Under the present publication policy, all technical papers so approved and recommended by the technical program committee are published in the AIEE *Transactions,* which is issued annually covering a calendar year and which serves as the permanent repository of material covering the progress and development of the electrical art. A *Transactions* section of the monthly magazine *Electrical Engineering* is set aside and allocated for publication of approved technical papers, this being in actuality a monthly preprint of identical pages which will constitute the current annual *Transactions* volume. Because of cost limitations, space has not been available in the monthly magazine for all approved papers. The overflow papers are printed twice yearly, in June and in December, in a special publication known as the "Supplement to *Electrical Engineering—Transactions* section." These supplements are available to Institute members at 50 cents each or $1 per year. Discussions of technical papers do not appear in *Electrical Engineering,* but first become available to the membership in the supplements, after a lapse of several months; for example, winter meeting discussions in the June supplement. Thus a member automatically receives most of the technical papers in the *Transactions* sections of *Electrical Engineering,* and by purchasing the supplements receives all of the remaining approved technical papers and all approved pertinent discussion. All this material appears in the annual AIEE *Transactions.* Discussions, however, do not appear on *Transactions* pages immediately adjoining the paper to which they apply, but appear elsewhere in the volume.

Certain of the papers submitted by contributors to the technical program committee and the technical committees for review and consideration are adjudged to be appropriate and timely for presentation and discussion at a national or District meeting, although they are deemed to be inappropriate for permanent record. Although these papers are presented at meetings, they are not published in the *Transactions* sections of *Electrical Engineering,* the supplements, or the *Transactions.* In current parlance they are known as "ACO" (advance copies only) papers; formally, this series of papers is designated as "AIEE miscellaneous papers" to distinguish them from the regular "AIEE technical papers" series.

To facilitate and to promote discussion of papers, individual pamphlet copies of each technical paper (regular or ACO) are issued in advance of the meetings at which the papers are scheduled to be presented and discussed. These pamphlets are known as "advance copies" or "preprints" of the technical papers. As the time element is of essential importance in relation to the meetings programs and speed is required in the physical reproduction of these papers to make them available to interested members as far as possible in advance of their presentation at meetings, a quick method of reproduction is resorted to that is known as the photolithographic or "photo-offset" process. In this process of reproduction, the author's original manuscript copy and illustrations are reproduced photographically approximately in facsimile, and without any of the intervening steps necessary for regular printing. The photolithographic process is a more expensive reproduction process than printing where substantial quantities are involved, but does offer the advantage of speed in reproduction. These "advance copies" or "preprints" therefore are produced only in limited quantities, and are made available to members approximately at cost.

To facilitate the efforts of District meeting committees in preparing programs best adapted to the timely and particular interests of the region to be served by District meeting, another class of paper has been given a recognized status by the Institute. This class of paper, known officially as "District papers" (DP) is authorized for acceptance for presentation and discussion at a District meeting solely at the discretion of the District meeting committee and without the requirement of advance review by the technical program committee or the technical committees. Correspondingly, such papers have status only as parts of the local District meetings programs, where they may be discussed along with other papers, but carry with them no commitment or obligation for publication of either

paper or discussion. In some instances, of course, the subject matter of District papers may be developed further and processed through the regular AIEE channels previously discussed and ultimately become eligible for inclusion in *Transactions*. These District papers normally are not prepared in pamphlet form by the Institute, nor are they generally available even at the meetings where they are scheduled for presentation, except in instances where authors may provide mimeographed or other copies.

Still another class of paper is involved currently in national and District meetings programs, known as "conference papers" (CP). These conference papers in a great many instances are not papers at all, but informal oral presentations made at conference sessions or round-table discussions which are organized from time to time on subjects of current interest under the encouragement of the technical program committee and the direct sponsorship of one or more of the technical committees. Like the previously mentioned District papers, conference papers are wholly informal presentations, not reproduced or distributed by the Institute, and carry no commitment for publication of paper or discussion. Such papers may be processed subsequently through regular channels and ultimately become *Transactions* material. Authors of conference papers sometimes make a limited number of mimeographed or other copies available at their own expense for use at the meetings. Otherwise, to preserve the complete informality of the conference session, there is no record of what transpires at such meetings or conferences, except as it may be covered in news reports in *Electrical Engineering*.

Electrical Engineering, under the present publication policy, contains many engineering and other articles, news items, personals, and other material of current general interest to AIEE members in addition to the *Transactions* section previously mentioned. This material is drawn from AIEE and other sources. The choice of material, as well as the general management and publication of the magazine, is the responsibility of the editor.

CURRENT PROBLEM

The electrical art has become so broad in scope and so specialized in nature that any particular technical program paper now is of interest to only a relatively small portion of the entire membership. Thus, under the present publication plan the average member is receiving many specialized papers in fields of relatively little interest to him. This repeatedly has been brought to the attention of the publication committee with the corollary suggestion that in most such instances an abstract or digest presenting on an informational basis the essence of such specialized papers would better satisfy the average member's requirement, leaving the full technical paper for the specialist. Under the present publication policy, the space required to print these technical papers in full in *Electrical Engineering* crowds out material of broader general interest, a situation that is growing more acute as the national and District meeting activities grow and the number of technical papers increases. The policy requiring publication of these papers in full in *Electrical Engineering* places undesirable restrictions and limitations upon the editor in his efforts, under his assigned responsibility, to produce in *Electrical Engineering* a magazine of greater service and more general appeal to the membership at large.

Discussions and rebuttals of technical papers, under the present publication plan, appear as already noted only in the *Transactions* and in the "Supplement to *Electrical Engineering—Transactions* section," and in any event only after the lapse of so long a time that much of the interest in the discussion is lost. Further, such discussions do not appear on pages immediately adjoining the papers to which they apply, thus necessitating reference to two different locations to obtain the complete story concerning any particular paper. The publication committee has received continual criticism of this practice.

NEW PUBLICATION PLAN

In the new publication plan, the *Transactions* section will be eliminated from *Electrical Engineering* and in its stead will be established a new series of publications to be known as "AIEE Proceedings" which will consist principally of all technical papers with their discussions, and which also may contain other appropriate material from national and District meetings. Individual members will be given the opportunity to select designated groups of these technical papers to be received free of charge, and other papers to be received as desired at a nominal charge to cover cost. More specifically the publications and policy of the Institute may be described as follows:

1. *Electrical Engineering*. Coincidental with the elimination of the *Transactions* section, the general-interest content of *Electrical Engineering* will be expanded, enhanced, and given greater flexibility to satisfy more fully the current and future requirements of the membership. It will be characterized especially by the following attributes.

(*a*). Subject only to general policy, the management and publication of *Electrical Engineering* will be entirely the responsibility of the editor.

(*b*). The general-interest content will be approximately doubled, contemplated to be at least 100 pages per month.

(*c*). The contents will be divided broadly into two general categories: (1) general articles, to be drawn from all currently available sources including material submitted for technical program papers after, but regardless of, action by the technical program committee, selected on the basis of timeliness and general-interest value to the membership at large and treated accordingly; (2) news material, comprehensively and appropriately reflecting all AIEE and related activities, also appropriately reflecting other current activities of interest and significance to the membership. For example, with reference to AIEE programs, whether technical sessions or conferences, it is contemplated that all technical program papers and all other appropriate material will be reflected currently in *Electrical Engineering* for the purpose of keeping the membership fully and currently informed.

(*d*). Summary abstracts of technical papers will continue to be included as at present and will be published as far as possible in advance of the meeting at which the papers are scheduled for presentation and discussion.

(*e*). The weight and quality of both inside and cover paper stock will be improved. This will recover the losses sustained incidental to United States Government wartime regulations and policies, and should obviate the justifiable criticisms that have been received from readers and advertisers concerning the unsatisfactory quality of reproduction of illustrations and other matter.

(*f*). A more aggressive advertising policy will be inaugurated to capitalize more fully upon the established prestige and substantial standing of *Electrical Engineering*.

(*g*). Distribution will continue to be made automatically to the entire AIEE membership; to be available also to nonmembers at an appropriate subscription price in accordance with postal regulations and as provided in AIEE bylaw 103 and 104.

2. *Preprints*. To encourage and facilitate discussion of technical papers at national and District meetings, the present practice of issuing advance pamphlet copies or "preprints," of technical program papers by the photolithographic process or an appropriate equivalent will be continued. Distribution of this material is to be made as follows.

(*a*). The minimum necessary number of copies will be available for distribution without charge by or at the specific and detailed request of the committee responsible for the development and approval of the manuscript in question. Such distribution is intended to be made prior to the meeting at which the paper is scheduled for presentation and discussion, and primarily for the purpose of stimulating qualified discussion.

(*b*). Copies will be available upon request to the entire AIEE membership at cost, such cost to include all production, overhead, and distribution costs; available to others at 100 per cent above members price. Following is the schedule of members prices:

Size of Pamphlet (Pages)	At Registration Desk	By Mail
12 or less	$0.10	$0.15
16	.15	.20
20	.20	.25
24 or more	.25	.30

3. *Proceedings*. For the purpose of better serving those among the membership who are especially interested in the full and

complete technical and mathematical details of subject matter within the field of their particular professional interest, a new series of publications is to be established, to be known as "AIEE *Proceedings*." The *Proceedings* will contain the full text of all papers approved by the technical program committee for publication in *Transactions* and for presentation at national or District meetings together with all discussions of those papers as approved by the technical program committee; also, may contain "miscellaneous (ACO) papers" approved by the technical program committee for presentation at national or District meetings; and other appropriate material. The *Proceedings* thus is planned to contain all material now published in the *Transactions* sections of the 12 monthly issues of *Electrical Engineering* and the two semiannual supplements, and in addition a very considerable amount of meeting material heretofore not published.

To provide maximum flexibility in production and distribution and thereby expedite materially the distribution of such material, and in deference to the long expressed wishes of the technical specialists among the Institute's membership who are most interested in the full and complete details of technical papers and their related discussions, the *Proceedings* will be issued in the form of individual printed pamphlets. Each pamphlet will contain the full text of an approved technical paper, and all approved pertinent discussion. The factors which will control the date of completion and initial distribution of this material will be (*a*) the determination by the technical program committee that no discussion is available or (*b*) the release by the technical program committee of discussion which it has approved for publication. On this basis, this material will be available to those who especially want it several months earlier than is possible under the present publication plan.

Each Fellow, Member, and Associate of the AIEE will be given an opportunity to select for receipt free of charge a limited quantity of the total number of *Proceedings* sections (pamphlets), the quantity designed to be sufficiently generous to enable the average member to obtain those papers in which he is most intimately concerned. Others will be available at cost. *Proceedings* sections also will be available at cost to Student Members and *Electrical Engineering* subscribers.

4. *Annual Transactions.* Current policies and procedures with reference to AIEE *Transactions* are to be continued without change, except that the flexibility introduced by the previously discussed changes in publication procedure will enable discussions to be assembled with the papers to which they apply. Bound volumes will continue to be issued on the basis of material presented during a calendar year, in single or multiple volumes, depending upon the amount of material and other considerations. *Transactions* will contain all formal AIEE technical papers and related discussion as approved by the AIEE technical program committee, and other record material appropriately approved. Distribution will continue on the basis of advance order; to the AIEE membership at cost, to others at appropriately higher prices.

5. *Special Publications.* To provide a facility in a field of publication services heretofore not satisfied adequately by the Institute, a series of "AIEE Special Publications" will be established and its use encouraged. It is contemplated that such publications will be issued from time to time and in appropriate relation to other AIEE publications to accommodate material of especial value and importance in its field but of limited general interest or applicability, or material peculiarly convenient and appropriate in the form of separate and substantially bound books or booklets. It is contemplated that these special publications may accommodate material highly technical or mathematical in content, or specialized in subject matter, such as the "AIEE Lightning Reference Book," the AIEE Joint Subcommittee report "Telemetering, Supervisory Control, and Associated Circuits," and the "Bibliography on Automatic Stations, 1930–1941," and other bibliographies, consolidated reprints, special reports, surveys, and similar material.

The format, typographical style, paper, and printing will be of a quality appropriate to an AIEE publication and commensurate with their positions in the AIEE family of publications. Special publications are to be issued on an entirely self-supporting basis, the proceeds from sales in effect constituting a revolving fund for continuation and stimulation of the project.

Distribution of special publications is to be on the basis of individual order, at cost to the AIEE membership; to others at 100 per cent above the prices which will be set for members.

6. *Reprint Service.* As a special service which has proved to meet many membership needs, the present practice of publishing quantities of reprints or preprints of AIEE published material will be continued. Such reprints usually will be produced on the basis of specific orders, but occasionally upon the basis of anticipated need. Such reprint services are to be performed on an entirely self-supporting basis.

7. *Annual Index.* To provide a consolidated single authoritative reference guide to all AIEE material published during any calendar year, a complete composite AIEE annual index is planned, to cover appropriately the content of *Electrical Engineering*, *Proceedings*, *Transactions*, and special publications. This index will continue the basic alphabetical multientry subject listing system originally developed by *Electrical Engineering* and used with eminent success over a period of many years. Distribution will be automatic and without charge to the AIEE membership and to regular subscribers to AIEE publications; upon request to others.

8. *Yearbook.* Publication of the Yearbook is to be continued as at present. Distribution is on the basis of individual requests.

9. *Conference Papers.* The informal AIEE conference papers will be treated very much as in the past. As such, they do not have publication status, although the editor of *Electrical Engineering* specifically is authorized to make appropriate use of conference material subject to the approval of the authors involved.

Great Lakes District Meets in Indianapolis

Sponsored jointly by the Fort Wayne and the Central Indiana Sections, the seventh Great Lakes District meeting was held in Indianapolis, Ind., October 9–11, 1946, with headquarters at the Claypool Hotel. Most of the sessions were held at the hotel the others were held in the auditorium of the Indiana War Memorial. The verified attendance was 298; other attendance details and comparisons with previous meetings of the Great Lakes District are given in accompanying tabulations.

The history of this meeting reveals quite a checkered career for a District meeting. Originally planned and scheduled to be held in Fort Wayne in April 1941, the meeting was indefinitely postponed in the war aftermath of Pearl Harbor. Picking up postwar where it had left off in 1941, the Fort Wayne Section had completed all plans and preliminary announcements to hold the meeting in Fort Wayne this fall, only to be overtaken by the hotel situation wherein Fort Wayne hotels withdrew the commitments of room and meeting space necessary. It was into this breach that the Central Indiana Section stepped, to provide the necessary hotel and other physical arrangements for the meeting in Indianapolis, and to join wholeheartedly with the Fort Wayne Section in carrying the meeting out to a very effective conclusion. Even the heavens collaborated, staging a once-in-a-century display shower of meteors on the opening night of the meeting.

CONFERENCE TYPE OF MEETING

A noteworthy feature was the fact that only four regular technical program papers were presented, the other 26 being informal presentations classed as "District papers" or "conference papers." This type of program proved to be both popular and effective in presenting for informal exchange and discussion a wide variety of up-to-the-minute topics which catered effectively to a very wide range of member interest. In fact, it drew a registered attendance of more than 80 nonmembers and many additional who did not register.

The 30 subjects were divided among six conference sessions covering electric machinery, electronics, power transmission and distribution, industrial power applications, communications, industrial control, and basic sciences. In general, the sessions ran two in parallel without too much conflict in member interest, the loss from conflict and interest being more than offset in the opinion of most observers by the additional scope of subject matter that could be accommodated by parallel sessions.

A more detailed summary report is given in another item elsewhere in these pages, covering the various technical topics discussed.

PROFESSIONAL ACTIVITIES DISCUSSED

One entire afternoon was devoted to a general conference on "Institute Activities and Study of the Organization of the Engineering Profession," under the chairmanship of AIEE Vice-President T. G. LeClair (F '40) of the Great Lakes District who also is chairman of the professional activities subcommittee of the AIEE committee on planning and co-ordination.

President J. Elmer Housley (F '43) discussed "How the Engineer Should Look at Tomorrow," stressing the need for engineers to give attention to problems of public and professional concern as well as to technical topics. He stressed the need for the AIEE to give steadily broader and better service to its membership, and briefly mentioned the efforts currently being made to accomplish this, and to learn more accurately what the membership wants by conducting discussions with representative groups throughout the country. He urged particularly much more active attention to industrial applications and problems, special efforts to develop the interest and greater participation of the executive and administrative group of electrical engineers, who deal primarily with management and labor relations problems rather than the more technical subjects; he urged especial effort to develop a better degree of co-ordination between the executive group and the younger engineers.

Vice-President LeClair presented a summary report covering the various projects that have been and are undergoing study by the committee on planning and co-ordination. As a preliminary to the specific discussion for which the session was planned originally, he also presented a comprehensive review of the study currently being made by his subcommittee on professional activities covering organization of the engineering profession, reflecting the high lights of the full report originally published in the April 1946 issue of *Electrical Engineering*, pages 169–73. Most of the discussions presented were from representatives of the various sections in the District, in response to previous solicitation. A few spontaneous comments were forthcoming from among the 12 persons in attendance including one from a visiting mechanical engineer who enthusiastically urged that "The engineers must organize as a profession or be organized by the government as a trade." In urging a single society for all engineers, one commentator made the broad assertion that "All the national societies are dominated by employers and hence are opposed to the over-all organization of individual engineers," but did not elaborate upon or substantiate this assertion. In general, the various discussers paid some homage to the theoretical idealism of the single-society idea, but as a matter of fact a substantial majority of all the commentators recommended "Plan C, Federation or Council of Existing Engineering Societies" and urged some early action in that direction.

Comparison of Great Lakes District Meeting Attendance

1946	October 9–11	Indianapolis, Ind.	298
1939	September 27–29	Minneapolis, Minn.	230
1935	October 24–25	West Lafayette, Ind.	481
1932	March 14–16	Milwaukee, Wis.	552
1929	December 2–4	Chicago, Ill.	600
1927	November 28–30	Chicago, Ill.	900
1926	May 6–7	Madison, Wis.	180

Analysis of Registration at Great Lakes District Meeting

Classification	Central Indiana Section	Fort Wayne Section	District 5*	Other Districts	Total
Members	68	19	74	31	192
Student Members	10				10
Men guests	50	13	15	5	83
Women guests	4	5	2	2	13
Totals	132	37	91	38	298

*Outside Central Indiana and Fort Wayne Sections.

ENTERTAINMENT AND TRIPS

The principal feature of the District meeting entertainment activities was the general banquet held Thursday evening, October 10, in the James Witcomb Riley room of the Claypool Hotel, with Vice-President T. C. LeClair presiding. Guests were treated to an excellent and lucid lecture on the "Nature and Implications of Atomic Energy," by Doctor Frank H. Spedding of Iowa State College, Ames. Doctor Spedding was in the midst of wartime atomic research from its earliest beginnings, and currently is a member of the Manhattan District Declassification Committee (MDDC), the group upon whose shoulders rests the responsibility of recommending those portions of the scientific information developed in wartime atomic research that can be released at this time to aid in scientific, technical, and medical research without impairing national security. While emphasizing the enormous potentialities of nuclear and atomic energy, for constructive use or for destruction, Doctor Spedding thoroughly debunked the sensational newspaper, magazine, and book accounts of quasi scientists and others which have caused many persons to fear that the earth itself was in dire danger of physical destruction. Doctor Spedding expressed the opinion that atomic power plants probably would become a reality within something like five years or so, especially for large naval and merchant ships, where the space required for fuel and power plants are of especially critical importance and economic significance.

To avoid interference with conference sessions, and at the same time to facilitate transportation and other arrangements, all technical inspection trips were scheduled for Friday afternoon after the last technical conference session. These trips provided for visits to the Marmon Herrington Company, manufacturers of mobile transporation equipment including all-electric trolley coaches; to the Esterline-Angus Company, Inc., makers of recording instruments; to the Eli Lilly and Company plant and laboratory, devoted to medicinal products; to various local power facilities of the Indianapolis Power and Light Company and of the Public Service Company of Indiana; to several of the plants of P. R. Mallory and Company, manufacturers of various radio controls and equipment; to the Chinook Mine of the Ayrshire Collieries Corporation at Staunton, Ind., and to the Maumee Collieries Corporation mine at Jasonville, Ind., two large strip-mining operations utilizing large electric shovel equipments; and to Purdue University, West Lafayette, Ind.

Activities especially aranged for women guests included a trip to the James Witcomb Riley Hospital for children, one of the outstanding hospitals for the medical care of children; a trip to the John Herron Art Institute; a tea and fashion show at the L. S. Ayres and Company, department store; the District meeting banquet.

Equipment exhibits originally had been planned in connection with the meeting, but did not materialize because of industrial production conditions.

COMMITTEES

The joint sponsorship of the Great Lakes District meeting in Indianapolis by the Fort Wayne Section and the Central Indiana Section is reflected in the committee organization. In general, Fort Wayne carried the responsibility for the development of the technical program and the Central Indiana Section carried the responsibility for arranging and managing the meeting activities. The personnel responsible for the success of the meeting include the following

General Cochairmen

F. H. Fleischer — S. C. Leibing

Meetings and Papers Committee

C. M. Summers, *chairman* — M. L. Schmidt
M. A. Baker — W. K. Self
F. O. Nottingham — E. S. Sullivan

Inspection Trips

C. E. Chatfield, *chairman* — W. A. Gentry
H. A. Berkheiser — C. E. Parks
L. E. Briscoe — C. R. Swenson
John Webb

Hotels, Registration, and Entertainment

S. C. Leibing, *chairman* — H. F. Roempke
L. T. Clipson — John Sears
E. G. Downie — Fred Stout
R. B. Hull — M. F. Schonefeld
G. W. Kintner — I. W. Strong
C. E. Parks — C. R. Swenson
J. R. Pies — R. O. Whitesell

Finance Committee

E. A. Linke, *chairman* W. H. Bollinger

Publicity Committee

J. L. Wright, *cochairman* S. A. Zimmerman, *cochairman*

Student Activities Committee

S. W. Winje *chairman*

Assistance in the actual handling of registration and related details was provided by the Indianapolis Convention and Visitors Bureau.

Contents Announced for December 1946 Supplement

Technical papers presented before AIEE technical meetings during 1946 and not published in the monthly *Transactions* section of *Electrical Engineering* or in the June supplement will appear in the December 1946 "Supplement to Electrical Engineering—Transactions Section." The December supplement will contain 30 technical papers, discussions of those papers, and discussions of the papers already published in the July–December monthly sections. Issuance of the December supplement will complete publication of the papers and discussions presented before the 1946 North Eastern District meeting in Buffalo, N. Y., April 24–25; the 1946 summer meeting, Detroit, Mich., June 24–28; the 1946 Pacific Coast meeting, Seattle, Wash., August 26–30; and the 1946 Great Lakes District meeting, Indianapolis, Ind., October 9–11.

In order that distribution of copies of the supplement can be made in February 1947, those desiring to receive copies should enter advance orders promptly.

Papers appearing in the December 1946 supplement, abstracts of which have been published in *Electrical Engineering* in advance of the meetings are:

Air Transportation

46-129—Electronic Control of Frequency Converter Sets for Testing Aircraft Models; *G. W. Heumann (A '44), W. F. Strong (A '41).* Abstracted in the June 1946 issue, page 278.

Basic Sciences

46-119—Formulas for Calculating the Inductance of Channels Located Back to Back; *W. G. Schwantz (Student), T. J. Higgins (M '46).* Abstracted in the June 1946 issue, page 279.

46-120—Negative Resistance Effects in Saturable Reactor Circuits; *J. M. Manley, E. Peterson (M '26).* Abstracted in the June 1946 issue, page 279.

46-195—The Use of Business Machines in Determining the Distribution of Load and Reactive Components on Power Line Networks; *P. D. Jennings, G. E. Quinan (F '18).* Abstracted in the August–September 1946 issue, page 411.

Communication

46-87—Radiotelemetering for Testing Aircraft in Flight; *C. L. Frederick (M '46).* Abstracted in the April 1946 issue, page 174.

46-183—Application of Compandors to Telephone Circuits; *C. W. Carter, Jr. (M '39), A. C. Dickieson (M '41), D. Mitchell.* Abstracted in the August–September 1946 issue, page 411.

Domestic and Commercial Applications

46-184—Specific Engineering Problems in Rural Electrification and Electroagriculture; *M. M. Samuels (F '24).* Abstracted in the August–September 1946 issue, page 412.

Electric Machinery

46-116—Electromechanical Transient Performance of Induction Motors; *C. N. Weygandt (A '37), S. Charp (A '42).* Abstracted in the June 1946 issue, page 280.

46-136—Current and Torque of D-C Machines on Short Circuit; *T. M. Linville (M '34).* Abstracted in the June 1946 issue, page 280.

46-164—Oscillations of a High-Voltage Secondary Winding; *A. Boyajian (F '26).* Abstracted in the June 1946 issue, page 280.

46165—Temperature Limits for Short-Time Overloads for Oil-Insulated Neutral Grounding Reactors and Transformers—II; *V. M. Montsinger (F '29), J. E. Clem (F '38).* Abstracted in the June 1946 issue, Page 281.

46-167—Rotating Electric Machine Time Constants at Low Speeds; *C. Concordia (M '37).* Abstracted in the June 1946 issue, page 281.

46-189—Dielectric Strength of Station and Line Insulation to Switching Surges; *P. L. Bellaschi (F '40), L. B. Rademacher (A '45).* Abstracted in the August–September issue, page 412.

46-202—Asynchronous and Single-Phase Operation of Synchronous Machines; *A. W. Rankin (M '45).* Abstracted in the October 1946 issue, page 472.

Induction and Dielectric Heating

46-124—Design of Induction-Heating Coils for Cylindrical Magnetic Loads; *J. T. Vaughan (A '44), J. W. Williamson (A '43).* Abstracted in the June 1946 issue, page 282.

Industrial Power Applications

46-194—Self-Excited Electromagnetic Drive for a Resonant Fatigue Machine; *A. R. Wilson (M '35).* Abstracted in the August–September 1946 issue, page 412.

Instruments and Measurements

46-133—Functional Analysis of Measurements; *I. F. Kinnard (F '43).* Abstracted in the June 1946 issue, page 283.

46-143—High-Vacuum Leak Testing With the Mass Spectrometer; *W. G. Worcester (A '42), E. G. Doughty.* Abstracted in the June 1946 issue, page 283.

46-181—An Oscillograph for Recording Transient Recovery Voltages; *W. G. Hoover (A '34).* Abstracted in the August–September 1946 issue, page 412.

Land Transportation

46-186—Comparisons of Railway Motive Power for Operations in the Pacific Northwest; *T. M. C. Martin (A '45).* Abstracted in the August–September 1946 issue, page 412.

Power Generation

46-108—Tennessee Valley Authority Hydroelectric Stations—Electrical and Mechanical Design; *R. A. Hopkins (F '43), H. J. Peterson.* Abstracted in the June 1946 issue, page 283.

46-144—Improvements in Performance of Hydroelectric Generating Units on the 2,000,000-Horsepower Saguenay System; *F. L. Lawton (M '36).* Abstracted in the June 1946 issue, page 284.

46-152—Modern Excitation Systems for Large Synchronous Machines; *J. B. McClure (A '29), S. I. Whittlesey (A '45), M. E. Hartmen.* Abstracted in the June 1946 issue, page 284.

46-154—The Development of Modern Excitation Systems for Synchronous Condensers and Generators; *F. M. Porter (M '45), J. H. Kinghorn (M '46).* Abstracted in the June 1946 issue, page 284.

46-182—Long-Distance Power Transmission as Influenced by Excitation Systems; *C. Concordia (M '37), S. B. Crary (F '45), F. J. Maginniss (A '43).* Abstracted in the August–September 1946 issue, page 413.

46-117—The Electrical Performance of Ceramic Dielectrics at Elevated Temperatures; *H. A. Frey (M '43), J. A. Jesatko (A '41).* Abstracted in the June 1946 issue, page 284.

Power Transmission and Distribution

46-196—Characteristics of a 400-Mile 230-Kv Series-Capacitor-Compensated Transmission System; *B. V. Hoard (M '36).* Abstracted in the August–September issue, page 413.

Protective Devices

46-170—Performance Criteria for Current Limiting Power Fuses—I; *E. W. Boehne (F '43), C. L. Schuck (A '39).* Abstracted in the June 1946 issue, page 285.

46-171—Performance Criteria for Current Limiting Power Fuses—II; *E. W. Boehne (F '43), C. L. Schuck (A '39).* Abstracted in the June 1946 issue, page 286.

46-173—Simplicity in Protective Relaying; *L. F. Hunt (F '38).* Abstracted in the June 1946 issue, page 286.

Future AIEE Meetings

Winter Meeting
New York, N. Y., January 27–31, 1947

North Eastern District Meeting
Worcester, Mass., April 23–25, 1947

Summer Meeting
Montreal, Quebec, Canada, June 9–13, 1947

Middle Eastern District Meeting
Dayton, Ohio, September 23–25, 1947

Midwest Meeting
Chicago, Ill., November 3–7, 1947

AIEE Survey on Vacuum Tubes Progresses

The survey of instrument manufacturers, in the interest of providing more reliable vacuum tubes for use in electronic instruments, is well under way, W. R. Clark (M '44) chairman of the joint subcommittee on electronic instruments, has announced.

Some of the items covered in the questionnaire, which has been mailed to more than 400 instrument manufacturers, are

1. Uniformity of tube characteristics.
2. Tube life.
3. Increased operation stability.
4. Reduction of microphonic effects.
5. Reduction of hum.
6. Lower noise level
7. Type of tube envelope.
8. Service conditions.
9. Yearly demand.
10. Cost of tubes.

The results of this survey will be tabulated and published as a committee report, and readers of *Electrical Engineering* are requested to assist in making it a comprehensive survey. Companies which have not received questionnaires may request them from the AIEE Joint Subcommittee on Electronic Instruments, AIEE Headquarters, 33 West 39th Street, New York 18, N. Y.

DISTRICT • • • • •

Southern District Students Convene in 1947

Tentative plans have been announced for a convention of the Student Branches of the Southern District to be held about the first week end in April 1947. The convention will open Thursday morning and extend through Saturday.

Convention headquarters will be on the campus of the University of Kentucky, Lexington. A number of student papers are expected to be submitted for prizes, and it is hoped that every Student Branch will be represented by at least one paper.

Included in the scheduled program are several speakers, a Thursday evening social, the reading of the student papers, tours of the bluegrass region, and a banquet and dance on Friday night. Special entertainment is planned for the women guests.

SECTION • • • • • •

Wire Strain Gauges Discussed by Canton Subsection

Professor K. F. Sibila (A '38) of the University of Akron (Ohio) was speaker at the first technical meeting of the Canton Subsection held October 14, on the premises of the Ohio Power Company.

Drawing on his experiences with the Guggenheim Institute during the war, Professor Sibila discussed the application of wire strain gauges to models in wind tunnels and the use of hot wires in the measurement of air velocities and directions.

ABSTRACTS • • • •

TECHNICAL PAPERS previewed in this section will be presented at the AIEE winter meeting, New York, N. Y., January 27–31, 1947, and will be distributed in advance pamphlet form as soon as they become available. Membe:s may obtain copies by mail from the AIEE order department, 33 West 39th Street, New York 18, N. Y., at prices indicated with the abstract; or at five cents less per copy if purchased at AIEE headquarters or at the meeting registration desk. Prices for nonmembers will be twice those for members.

Mail orders will be filled AS PAMPHLETS BECOME AVAILABLE

Air Transportation

47-5—Parallel Operation of Aircraft Alternators Using Electronic Frequency Changers; *O. W. Bowlus (M '46), P. T. Nims (A '40). 25 cents.* This paper reports initial work on an electronic method for operating several main aircraft engine

Attending the banquet held September 21, 1946, to celebrate the opening of the new headquarters of the Engineers' Society of Milwaukee were AIEE President J. Elmer Housley (F '43) and the following members of the Milwaukee Section: (front row, left to right) L. M. Swenson (A '41) Allis-Chalmers Manufacturing Company, chairman of the Section program committee; F. J. Van Zeeland (M '44) professor at the Milwaukee School of Engineering, and secretary–treasurer of the Section; Mr. Housley; K. L. Hansen (F '34) Harnischfeger Corporation, past chairman of the Section; E. L. McClure (M '41) Wisconsin Electric Power Company, director of the Section; L. C. Aicher, Jr. (M '43) Allis-Chalmers Manufacturing Company, director of the Section. In the second row are: C. P. Feldhausen (M '40) Cutler-Hammer, Inc., director of the Section; S. H. Mortensen (F '20) Allis-Chalmers Manufacturing Company, past Section chairman; W. C. Johnson, Allis-Chalmers Manufacturing Company; Walther Richter (F '42) Allis-Chalmers Manufacturing Company, Section chairman; E. J. Limpel (M '44) A. O. Smith Corporation, director of the Section; E. T. Sherwood (M '44) Globe-Union, Inc., director of the Section; H. C. Brem (A '43) Allis-Chalmers Manufacturing Company, director of the Section

driven alternators in parallel to supply the electrical loads of an airplane. This would make unnecessary any mechanical constant speed drive for each generator. An electronic frequency changer is associated with each alternator and the outputs of the changers paralleled at the load bus. Synchronization, load division, and provision for reactive power are discussed, and test data obtained from two 4-kva units are presented.

47-11—The Calibration of Ignition Crest Voltmeters; *W. L. Davis, C. E. Warren. 15 cents.* Field experience with crest voltmeters for use on internal combustion engine ignition systems has shown variation in readings between several types of meters connected to the same point in an ignition system. This paper outlines a study of the factors influencing the performance of several types of meters. Apparatus is described for producing repetitive voltage surges the time of which to crest, shape of tail, amplitude, and repetition rate are easily adjustable over a considerable range. The equipment was used to determine the effects of the variables mentioned on the operation of several types of ignition crest voltmeters. A high-speed cathode-ray oscillograph was used to calibrate the meters on various types of voltage surges. Typical results for two meters are given in the form of calibration areas. Effects of voltage wave shape on the meter readings are discussed.

Basic Sciences

47-1—Eddy Currents in Disks: Driving and Damping Forces and Torques; *A. D. Moore (F '43). 30 cents.* Driving torque due to a-c quadrature fluxes through a disk can be solved for by means of theory and procedures presented. Fluxes may have any distribution. A facilitating chart is described. As an example, the driving torque of a watt-hour meter is solved for. Damping torque of a disk rotating through constant fluxes can be solved for by means of theory and procedures developed from what may be a new concept. Flux areas may be irregular. The methods can be extended to cover cases of nonuniform density. A facilitating chart is described. As an example, the damping torque of a watt-hour meter is solved for. Damping of disks moved translationally is covered by a family of curves, by which damping for accelerometers and other devices can be designed in direct fashion. For both driving and damping torques, reciprocity laws are proved.

47-2—Theory and Application of Complex Logarithmic and Geometric Mean Distances; *T. J. Higgins (M '46). 20 cents.* Logarithmic and geometric mean distance theory is utilized both in the design and in the calculation of the resulting electrical and mechanical performance of power and transmission lines, industrial distribution busses, transformer windings, coils and solenoids for various purposes, and yet other current-carrying apparatus. Heretofore calculation of the required logarithmic and geometric mean distances has been effected by direct integration of the multiple integral,

$$\int_{S_2}\int_{S_1} \log\,[(X-x)^2+(Y-y)^2]^{1/2}dS_2dS_2$$

wherein (X, Y) and (x, y) indicate arbitrary points in the lineal or areal cross sections S_1 and S_2 of the current-carrying apparatus. Except in certain simple well-known cases, evaluation of this multiple integral requires very heavy analysis. This difficulty is obviated by use of the theory advanced in this paper. By use of simple conformal mapping, the foregoing real integral is expressed as the real or imaginary part of a complex integral which easily is evaluated. The resulting complex logarithmic mean distance comprises as its readily recognized real or imaginary part the desired real logarithmic and geometric mean distances. Calculation of several logarithmic mean distances typical of those used in practice illustrates application of the general theory and the advantages to be gained by use of it.

Electric Machinery

47-3—Equivalent Circuit of the Primitive Rotating Machine With Asymmetrical Stator and Rotor; *Gabriel Kron (M '45). 20 cents.* The equations and equivalent circuits of all standard types of rotating electric machinery operating under arbitrary conditions may be derived as special cases of a single generalized rotating machine, the so-called primitive machine. This paper derives the steady-state equivalent circuit of the primitive machine having *unbalanced rotor* as well as stator. That is, the load or transmission line connected to a salient-pole synchronous machine is assumed to be unbalanced (due to faults, short circuits, and the like) or the load connected to the slip rings of an induction motor is assumed to be unbalanced. The single-phase alternator, instrument Selsyns and so forth, as well as sudden short-circuit studies are cases that require an infinite series of time harmonics in the stator and rotor windings in addition to the fundamental. It is assumed that these time harmonics are due to space waves with fundamental pairs of poles rotating at different speeds. The effect of the harmonic space waves is not considered. A companion paper by Rankin shows network analyzer studies made of the harmonic currents and torques of a single-phase alternator.

47-9—The Dielectric Properties of Cellulow Insulation Impregnated With Semiconductor Liquids; *F. M. Clark (A '24). 20 cents.* Liquid impregnants for cellulosic insulation are classed as dielectric, semiconducting, and conducting. The low power factor of insulation impregnated with a dielectric liquid rises to high value as the liquid resistivity decreases from the dielectric to the semiconducting classification. In the semiconducting grouping, there is a limited range of resistivity which is accompanied by a marked drop in the power factor of the treated insulation. With kraft paper, the power factor falls to about three per cent. Because of the conducting nature of the liquid, the effective dielectric constant obtained simultaneously with such low power factor is abnormally high. The Permalytic capacitor is a kraft paper spaced capacitor of the usual type, impregnated with a semiconducting liquid. This capacitor is characterized by a physical size equal to about 30 per cent of the physical volume per kilovolt-ampere of the corresponding oil-treated capacitor. The power factor of the Permalytic capacitor is about three per cent. This type of capacitor offers interesting possibilities for those applications where a small dependable low-cost low-voltage capacitor is required.

47-10—Simplified Graphical Method of Computing Thermal Transients; *Paul Narbutovskih (M '43). 15 cents.* During recent years the subject of estimating the transient overload capacity of electric equipment, transformers in particular, has gained considerable importance among operators and designers. Practically acceptable methods of making such estimates have been developed, including graphs and charts, which simplify the computations of certain specific problems involved. This paper gives a new graphical method of computing thermal transients which possesses a considerable degree of simplicity and flexibility. The method is based on the use of new, here proposed, transient co-ordinate paper on which any single-capacity thermal or electrical exponential transient will plot as a straight line. The use of this paper greatly simplifies the solution of an exponential equation for temperature and time. It adapts itself readily to handling transient temperatures in case of complex load cycles, as well as more intricate problems involved in equipment design.

Instruments and Measurements

47-12—A Comparison of Two Basic Servomechanism Types; *Herbert Harris (A '46). 30 cents.* Servomechanisms with lead network stabilization and those with speed feedback through a high pass filter are analyzed and compared. The method of analysis is on a frequency response basis using inverse transfer functions. Detailed design tabulations are given along with the limitations imposed by response delays. Analysis of motor suitability for these servomechanisms shows that torque-to-inertia ratio is a poor quantity for a figure of merit, and several alternative ratios are shown to have more value.

47-7—The Application of Lead Networks and Sinusoidal Analysis to Automatic Control Systems; *G. J. Schwartz (A '46). 30 cents.* This paper discusses the general design considerations relating to the construction of a high performance servo-

mechanism. The sinusoidal method of analysis is applied, and transfer loci for complete servo systems and their individual components are presented in order to illustrate the theoretical results obtained. Particular emphasis is placed upon the effect of anticipation or lead networks on the performance of a servomechanism. Direct reference to an existing armature-controlled d-c motor servomechanism is made throughout, and the correlation between theory and practice as applied to a servomechanism design is demonstrated.

47-13—The Cathode-Ray Spectrograph; *Rudolf Feldt (M '46), Carl Berkley. 15 cents.* The instrument described has been developed by the Allen B. DuMont Laboratories Inc., and produces complete spectral distribution curves of the source under test at the rate of 240 complete spectra per second. Spectrogram is presented on a cathode-ray tube. The spectrograph consists of four units which may be used independently. These are: source, capacitating system, scanning means, and indicator. Wave length calibration is done by replacing source with a gaseous discharge lamp. Sensitivity is adjusted so that resultant curve represents visual response. Dispersion is given by prism with scanning effected by a synchronous, vibrating, resonant mirror at 120 cycles per second. A suitably filtered multiplier photoelectric cell is used. A standard cathode-ray oscillograph may be used as indicator, but special d-c oscillographs will be available for this purpose. Samples may be compared by switching or use of delay screens. This instrument was developed initially to study the color of microsecond current pulses on a cathode-ray tube fluorescent screen. Results are shown. The color of any other varying light source can be studied similarly. Provision is made for insertion of transmission samples, chemical samples, or filters for calibration. Continuous recording is possible with camera attachment. Indication may be remote or separated from analyzer. Possible applications are:

1. Instantaneous color matching.
2. Fluorescent lamp characteristics.
3. Spectrographic research.
4. Continuous process control.
5. Transmission of spectral characteristics from remote or inaccessible locations.

Power Generation

47-4—Theoretical Approach to Speed and Tie Line Control; *Robert Brandt (A '26). 15 cents.* Frequency error is an indicator of the amount of excess or deficiency of generation on a whole interconnection, and a frequency-generation line may be established for the whole or for any section which is separated from all others by tie lines over which control may be desired. Automatic equipment is available, using these fundamentals under the name of load biased frequency control, which will make the required adjustments to generation to restore normal speed and tie line loadings with a minimum of false moves. By this means each section will handle, in normal operation, only its own load changes, regardless of the speed of response of its prime movers to correcting impulses. Methods of automatic control by the selective-frequency blocking system and the master flat-frequency station method are examined and found wanting. The use of uncontrolled speed-sensitive governors with small droop is shown to be disadvantageous, if operated in parallel with automatic bias equipment. Consideration is given to practical matters of operation, including an improvement in handling manual generation shifts, action of controller in time of trouble, effect of errors in determining the frequency-generation line, and desirable characteristics of governors.

Land Transportation

47-8—Developments in Current Collectors for High Speed Service; *B. F. Langer. 15 cents.* Theoretical studies and wind-tunnel tests have been used to develop an improved pantograph shoe and shoe mounting. The shoe itself has been made aerodynamically stable, and the supporting springs make it follow the wire smoothly at all speeds.

Protective Devices

47-6—Resistors for 138-Kv Cable Switching; *E. K. Sadler (M '29), T. M. Blakeslee (M '38). 25 cents.* Installations of 138-kv underground cable by the Los Angeles Department of Water and Power provided a field of laboratory for capacitive-current switching tests. This paper describes and presents results of an investigation of overvoltages incident to switching the charging current of these cable circuits. The development and performance of step-type resistors installed in 138-kv circuit breakers for limiting these overvoltages are described. A theoretical discussion of the effect of resistors in circuit breakers handling capacitive currents is given in an appendix.

PERSONAL • • • • •

Lewis Warrington Chubb (A '09, F '21) director of research, research laboratories, Westinghouse Electric Corporation, East Pittsburgh, Pa., has been awarded the John Fritz Medal for 1947 for "pioneering genius and notable achievements during a long career devoted to the scientific advancement of the production and utilization of electric energy." Doctor Chubb was born October 22, 1882, in Fort Yates, N. Dak., and was graduated from Ohio State University in 1905 with the degree of mechanical engineer in electrical engineering. He became an engineer apprentice with the Westinghouse corporation and from 1906 to 1910 worked on the development and testing of sheet

L. W. Chubb

steel and did research on transformers. He was placed in charge of the electrotechnical and magnetic research section in 1910 and of the research work of the material and process engineering department in 1919. When the radio engineering department was organized in 1920, Doctor Chubb was made manager. He left East Pittsburgh in 1930 when the radio activities of the Westinghouse corporation, the General Electric Company, and the Radio Corporation of America were consolidated at Camden, N. J., to become assistant to the vice-president of RCA. Later that year he returned to East Pittsburgh as director of research. Doctor Chubb's contribution to the knowledge of magnetic properties of iron and iron alloys, and his improvements in the design of electric machinery and in the measurements of electrical and magnetic quantities have had great influence in the development of electrical engineering during the past 30 years. He has been granted about 200 patents for electrical, mechanical, chemical, electrochemical and instrumental developments. He has received the degree of doctor of science from Allegheny College and from the University of Pittsburgh. In 1934 he was awarded the Lamme Medal of Ohio State University. Author of many technical articles, he is a member of the American Association for the Advancement of Science, the American Physical Society, the Franklin Institute, the Institute of Radio Engineers, the Illuminating Engineering Society, and the National Electrical Manufacturers Association. He also is a member of Tau Beta Pi and Sigma Xi. He formerly was a member of the International Electrotechnical Commission.

C. A. Robinson (A '11, F '22) vice-president, secretary, and treasurer, Chesapeake and Potomac Telephone Company, Washing, D. C., has retired. Mr. Robinson received the degree of bachelor of arts from Johns Hopkins University in 1903 and the degree of mechanical engineer from Cornell University in 1906. Immediately afterwards he entered the engineering department of the New York Telephone Company, New York. From 1909 to 1919 he was employed in the telephone transmission branch of the engineering department

of the American Telephone and Telegraph Company, New York. In 1919 he was sent by the Western Electric Company to Europe in connection with engineering a large toll telephone cable project in France. He was appointed transmission engineer of the American Telephone company later that year and in 1920 became chief engineer of the Chesapeake company. He was appointed assistant vice-president in 1929 and general manager of the West Virginia company in 1931. He was elected vice-president and general manager of the Washington company in 1940.

L. A. Kilgore (A '29, M '37) formerly head of the rectifier and motor section of the a-c generator department, Westinghouse Electric Corporation, East Pittsburgh, Pa., has been appointed assistant manager of the a-c generator department. Mr. Kilgore joined the corporation in 1927 after his graduation from the University of Nebraska. Mr. Kilgore is an ex-chairman of the AIEE committee on electric machinery and a member of Sigma Xi. **M. R. Lory** (A '40) has been named manager of the motor section of the department and **J. L. Boyer** (A '43) manager of rectifier development. Mr. Lory, who holds the degree of bachelor of science from Colorado State College and the degree of master of science from Massachusetts Institute of Technology, joined the Westinghouse corporation in 1928. Mr. Boyer has been with the corporation since he was graduated from Rice Institute in 1942. He is a member of Tau Beta Pi.

L. R. Milburn (A '20, F '39) electrical engineer, Great Lakes Steel Corporation, Ecorse, Mich., has been elected president of the Association of Iron and Steel Engineers. Mr. Milburn was graduated from the University of Michigan in 1918, and, after service in the United States Army, he entered the test course of the General Electric Company at Schenectady, N. Y., and Pittsfield, Mass. He was transferred to the Cleveland, Ohio, office in 1919 as complaint and installation engineer and to the Cincinnati, Ohio, office in 1923. From 1925 to 1929 he was engaged in construction work for the Louisville Gas and Electric Company. He joined the Great Lakes company as electrical draftsman in 1929, became acting electrical engineer in 1933 and electrical engineer in 1935.

P. H. Maurer (A '33) formerly commercial application engineer, National Pneumatic Company, New York, N. Y., has been appointed chief engineer of the company. Mr. Maurer attended Newark College of Engineering and the University of Michigan. He previously was associated with the Bendix Aviation Corporation and with the Hudson Motor Car Company as electrical engineer in charge of car engineering. At the outbreak of the war he was loaned to Sparks-Withington Company, Jackson, Mich., which was building link trainers but later he returned to the Hudson company which was operating the Naval Ordnance Plant, Centerline, Mich. He was appointed executive engineer of the Redmond Company, Owosso, Mich., and in 1945 joined the National Pneumatic Company. He is a member of the Engineering Society of Detroit and the Society of Automotive Engineers.

E. T. Augustine (A '27, M '33) formerly assistant chief engineer, Western Massachusetts Electric Company, Springfield, has been named assistant general superintendent of the company. Mr. Augustine was employed by the company in 1929 as system planning engineer. He was district engineer from 1932 to 1944 in which year he was appointed assistant chief engineer. **M. G. Moses** (A '41, M '42) formerly planning engineer, has been appointed assistant chief engineer of the system. A 1927 graduate of the University of Minnesota, he was employed in the radio research laboratory at Harvard University, Cambridge, Mass., from 1943 to 1945, the year he joined the Western Massachusetts company. Previously he had been associated with the Northwestern Public Service Company and the Northern States Power Company.

A. B. Campbell (A '20, F '38) formerly eastern representative of Hughes Brothers, New York, N. Y., has been appointed executive secretary of the National Association of Corrosion Engineers with headquarters in Houston, Tex. Mr. Campbell holds the degrees of bachelor of science and electrical engineer from the University of Illinois. He joined the faculty of Iowa State College in 1915 and was appointed to the Iowa Board of Railroad Commissioners in 1920. He became associated with the National Electric Light Association in 1924 and continued with its successor, the Edison Electric Institute, until 1943, at which time he joined Hughes Brothers.

D. C. Inman (A '38, M '45) formerly engineering and service manager for the Westinghouse Electric Corporation, Cleveland, Ohio, has been appointed engineering manager of the district engineering and service department, East Pittsburgh, Pa. A graduate of Colorado Agricultural College, Mr. Inman joined the company in 1925. After ten years in the control engineering department, he joined the Pittsburgh, Pa., engineering division of the company as consulting and application engineer. In 1938 he was transferred to the Cleveland office and became manager in 1941. He is a past president of the Cleveland Engineering Society.

Hector Sleeman (M '27, F '36) colonel in the British Army, who during the war served in Burma and India with the Royal Artillery and the Indian Electrical and Mechanical Engineers, will be released from active service in January 1947. After a trip to engineering establishments in Great Britain, Canada, and the United States, Colonel Sleeman intends opening a consulting firm in Melbourne, Australia. Before entering military service in 1939, Colonel Sleeman was chief engineer of the Rangoon (Burma) Electric Tramways and Supply Company.

J. O. Johnson (A '36, M '43) formerly chief product engineer, product engineering division, Aircraft-Marine Products, Inc., Harrisburg, Pa., has been appointed assistant general manager of the Buchanan Electrical Products Company, Inc. Mr. Johnson, who was graduated from Rensselaer Polytechnic Institute in 1931, was associated with Sears Roebuck and Company, Chicago, Ill., until 1933. He was assistant to the chief electrical engineer at the Material Laboratory, United States Navy Department, Brooklyn, N. Y., from 1934 to 1940, and became chief product engineer with Aircraft-Marine Products in 1941.

W. D. Coolidge (A '10, M '34) retired director of the General Electric Research Laboratory, Schenectady, N. Y., has been appointed to head the new General Electric atomic research and development laboratory at the Hanford engineer works near Richland, Wash. Doctor Coolidge has been X-ray consultant to the General Electric Company since his retirement in 1945 (*EE, Mar '45, p 125*). He has been associated with the atomic bomb project since 1941 at which time he was one of the six men selected to evaluate the military importance of uranium.

W. A. MacCrehan, Jr. (A '43) formerly supervisor of quality control in the motor division, General Electric Company, Lynn, Mass., has been appointed assistant professor of administrative engineering and director of the Gage Laboratory at New York University, New York, N. Y. Mr. MacCrehan has been instructor for the State University extension courses in quality control at Harvard University, Cambridge, for the past three years and at Boston University Evening College of Commerce for 1945–46. He is a member of the National Society for Quality Control, the American Standards Association, and the American Society of Tool Engineers.

S. G. Lutz (A '38) formerly head of the measurements and direction-finding section of the Naval Research Laboratory, Washington, D. C., has been appointed chairman of the electrical engineering department of the College of Engineering of New York (N. Y.) University. Doctor Lutz received from Purdue University the degree of bachelor of science in 1933, master of science in 1934 and doctor of philosophy in 1938. He was assistant professor at Southern Methodist University, Dallas, Tex., from 1938 to 1940. In 1940 he joined the Naval Laboratory. He is a member of the Institute of Radio Engineers and the American Society for Engineering Education.

G. L. Rosenberger (A '31) superintendent of construction, Virginia Electric and Power Company, Alexandria, has retired. Mr. Rosenberger has been with the Virginia Power company and its predecessors since 1924. Previously he was superintendent of public utilities for Manassas, Va., from 1914 to 1920 and superintendent of the Herndon Light and Supply Company from 1920 to 1922. As superintendent of the Fairfax–Loudon Light and Power Company from 1922 to 1924, he built the first transmission line between Alexandria and Leesburg, Va.

W. G. Worcester (A '42) engineer, general engineering laboratory, General Electric Company, Schenectady, N. Y., has been appointed assistant professor of electrical engineering at the University of Colorado, Boulder. Professor Worcester was graduated from the University of Colorado in 1939 with the degree of bachelor of science and received the degree of master of science from California Institute of Technology in 1940. He has been with the General Electric Company since 1941. He is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

E. R. McKee (A '30, M '36) professor and head of the department of electrical engineering, University of Vermont, Burlington, and acting dean of the college of engineering, has been named dean of the new College of Technology at the University. Professor McKee, who holds the degrees of bachelor of science, master of science, and electrical engineer from Iowa State College, has been a member of the faculty of the University of Vermont since 1934.

T. J. Bostwick (A '18, M '26) chief electrical engineer, Aluminum Company of America, Pittsburgh, Pa., has retired. However Mr. Bostwick will remain available to the company and its subsidiaries as consulting engineer. Mr. Bostwick was born in 1876 in Oswego, N. Y. After two years in the students' course at the Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pa., and three as electrician for the Niagara Falls (N. Y.) Power Company, he joined the Aluminum Company in 1900 as electrical engineer. In 1912 he was made assistant chief electrical engineer and in 1924 chief electrical engineer.

H. E. Kent (A '25, M '31) engineer with the Edison Electric Institute, New York, N. Y., has been appointed director of engineering. Mr. Kent, who was born in 1901 in Providence, R. I., holds the degrees of bachelor and master of science from Massachusetts Institute of Technology. Mr. Kent has been associated first with the National Electric Light Association and then with the Edison Electric Institute since 1924. He has been concerned particularly with the problem of the inductive co-ordination of power and communication circuits. He is a member of Tau Beta Pi.

F. E. Nutt (A '36) formerly assistant sales manager, Crocker-Wheeler Electric Manufacturing Company, division of Joshua Hendy Iron Works, Ampere, N. J., has been appointed executive engineer of the division. Mr. Nutt is a 1931 graduate of the University of Denver. Before joining the Crocker-Wheeler sales department in 1942 he was a student engineer and assistant in the high voltage laboratory of the General Electric Company, Pittsfield, Mass. He was Crocker-Wheeler sales representative in Washington, D. C., from 1943 until in 1945 he was made assistant sales manager.

C. S. Allen (A '31, M '37) formerly assistant general manager, Sawyer Electrical Manufacturing Company of Los Angeles, Calif., has been elected vice-president and general manager of the company. Mr. Allen was graduated from Mississippi State College in 1929 and, after completing the test course at the General Electric Company, was assigned to the fractional horsepower motor engineering department as design engineer. He joined the Sawyer Company as chief engineer in 1945.

R. C. Mason (A '26, M '37) of the electrophysics department, Westinghouse Research Laboratories, East Pittsburgh, Pa.; **J. W. Simpson** (A '43) section engineer of the company's switchboard engineering department; and **A. H. Toepfer** (A '31, M '39) application engineer in the district engineering and service department; have been loaned for one year to the Institute of Nuclear Studies which is being established at Oak Ridge, Tenn., to provide channels for co-operative research between government, universities, and industrial concerns.

E. I. Smith (A '32) until recently with the Signal Corps, has been appointed telegraph superintendent of the Atlantic Coast Line Railroad Company, New York, N. Y. Mr. Smith was with the Western Union Telegraph Company from 1931 until he entered military service in 1942. During the war he served as chief signal officer, radio station WAR, Washington, D. C., and spent 34 months in Africa and Europe.

R. W. Harper (A '26, M '36) of Bell Telephone Laboratories, Inc., New York, N. Y., and recently lieutenant colonel and commanding officer of the 71st Region, Plant and Engineering Agency, Army Communications Service, has received the Army Commendation Ribbon. The citation states that he "capably discharged important responsibilities in directing and co-ordinating the construction and installation of radio communication facilities and navigational aids for aircraft to complete communications along the routes of the Air Transport Command."

A. W. Post (A '06, M '28) equipment cost studies engineer, long lines department, American Telephone and Telegraph Company, New York, N. Y., has retired. Mr. Post was graduated from the Polytechnic Institute of Brooklyn in 1909 and that year entered the plant department of the long lines department. He was transferred to the engineering department in 1917. In recent years he has been concerned with costs of central office and other equipment associated with the long lines plant.

H. C. Purcell (A '41) formerly electrical designer, engineering department, Hawaiian Howard Electric Company, Honolulu, T. H., and **C. R. Day** (A '44) formerly assistant engineer, Modesto (Calif.) Irrigation District, have been appointed assistant designing engineers for the Sacramento (Calif.) Municipal Utility District. Mr. Purcell is a graduate of Stanford University and Mr. Day of the University of California.

D. H. Lauder (M '45) formerly assistant manager, service engineering division, General Electric Company, Schenectady, N. Y., has been appointed works manager of the company's newly acquired Hanford (Wash.) engineer works (*EE, July '46, p 361*). The atomic energy research in progress at Hanford has been designated the General Electric Nucleonics Project. Mr. Lauder has been with the company since he was graduated from the University of Pittsburgh in 1922.

A. B. Morgan (A '28, M '35) formerly rate and power consultant, Edison Electric Institute, New York, N. Y., has been appointed managing director of the institute. A graduate of Massachusetts Institute of Technology, Mr. Morgan joined the National Electric Light Association, predecessor of EEI, in 1927 as an assistant engineer. He was titled engineer and rate and power consultant in 1933.

A. W. Melloh (A '38, M '45) senior telephone engineer, Stromberg-Carlson Company, Rochester, N. Y., has been appointed to the telephone transmission group and will be responsible for the development of carrier-current equipment. Mr. Melloh, a graduate of the University of Minnesota, trained officers and enlisted personnel in the operation and maintenance of underwater sound equipment at the Navy's San Diego (Calif.) laboratory during the war.

J. W. Mullally (A '45) formerly manager of bare and weatherproof sales for the Anaconda Wire and Cable Company, New York, N. Y., has been appointed manager of utility sales. Previously Mr. Mullally served as district manager for the company in Pittsburgh, Pa. During the war he served for three years as chief of the Wire Mill Branch of the Copper Division of the War Production Board.

D. I. Bohn (A '23, M '41) formerly electrical engineer, has been appointed chief electrical engineer of the Aluminum Company of America, Pittsburgh, Pa. Mr. Bohn, who was graduated from the University of Wisconsin in 1921, entered the employ of the company as assistant electrical superintendent in Badin, N. C., in

1923 and was made superintendent in 1925. He was transferred to Pittsburgh as electrical engineer in 1928.

E. A. Ossmann (A '44) electronic engineer, Allen B. Du Mont Laboratories, Inc., has been named New York State factory representative for the company. Mr. Ossmann will have his headquarters in Rochester, N. Y. A graduate of Manhattan College, Mr. Ossmann previously was associated with the Anaconda Copper Corporation and the General Electric Company and served in the Armed Forces.

G. P. Lehmann (A '35) formerly colonel, 12th Air Force Service Command, has been appointed assistant manager of the plastics division, General Electric Company, Pittsfield, Mass. Mr. Lehmann had been with the General Electric Company from 1935 until he went on active duty as a lieutenant in 1941.

J. H. Howard (A '41) research supervisor, Engineering Research Associates, St. Paul, Minn., has received the award of the Legion of Merit from the United States Navy in recognition of outstanding service performed while on duty with the Division of Naval Communications. Mr. Howard served in the Navy from 1943 to 1946 and was a lieutenant at the time of his release from duty.

F. B. Silsbee (A '13, F '42) formerly chief of the electric instruments section, National Bureau of Standards, Washington, D. C., has been appointed chief of the division of electricity of the bureau. Doctor Silsbee, who holds the degrees of bachelor and master of science from Massachusetts Institute of Technology and received the degree of doctor of philosophy from Harvard University in 1915, has been with the National Bureau of Standards since 1911. He became chief of the electric instruments section in 1939.

W. R. Lyon (A '20, M '26) recently lieutenant colonel in the Army Air Forces, Wright Field, Ohio, has returned to Bell Telephones Laboratories, Inc., New York, N. Y. For about half of his five and a half years in the Army, he was resident representative in a number of manufacturing plants in Ohio and Indiana. For over two years he was successively instructor in electrical engineering at the AAF Engineering School and professor at the AAF Institute of Technology, Wright Field.

D. E. B. Corson (A '33, M '41) formerly sales manager of the power division of Aerovox Corporation, New Bedford, Mass., has been appointed manager of the special products division of Solar Manufacturing Corporation, New York, N. Y. A 1932 graduate of Massachusetts Institute of Technology, Mr. Corson previously was sales manager of the power factor division of Cornell-Dublier Electric Corporation, South Plainfield, N. J.

Mary F. Blade (A '35) who has been appointed instructor in machine drawing at Cooper Union, New York, N. Y., will be the first woman faculty member in the school of engineering in 87 years. Mrs. Blade, who was graduated from the University of Utah in 1934, since has been electrical engineer for the Southern Utah Power Company, Cedar City, and methods engineer for the United States Employment Service.

L. W. Whitton (A '40) formerly production manager, Otis Elevator Company, New York, N. Y., has been appointed manager of operations. A 1922 graduate of the University of California, Mr. Whitton joined the company in 1919 in San Francisco, Calif., was transferred to New York in 1936, and has been production manager since 1944.

W. E. Tinsley (A '44) works engineer, Sperry Gyroscope Company, Great Neck, N. Y., is one of the first three engineers to receive graduate scholarships under a newly instituted plan of the company. Mr. Tinsley is studying for the degree of master of science in aeronautical engineering at New York University.

D. A. Quarles (A '23, F '41) director of apparatus development, Bell Telephone Laboratories, Inc., New York, N. Y., has become a member of the committee on electronics of the Joint Research Board. Three officers each from the Army and Navy and two other civilians complete the committee.

E. E. Browning, Jr. (M '38) formerly plant extension engineer, operations and engineering department, American Telephone and Telegraph Company, New York, N. Y., has been placed in charge of a new section on engineering results in that department with the title of results engineer.

G. W. Morgan (A '37) formerly design electrical engineer, Gulf States Utilities Company, Beaumont, Tex., has joined the Hawaiian Electric Company, Ltd., Honolulu. Mr. Morgan, who joined the Gulf States company in 1940, previously had worked for Stone and Webster Engineering Corporation.

E. M. Strong (A '26, M '40) professor and chairman of the electrical engineering division, Cornell University, Ithaca, N. Y., has been appointed to represent the Illuminating Engineering Society in the division of engineering and industrial research of the National Research Council.

Edward Kerschner (A '19) formerly eastern district manager, General Cable Corporation, New York, N. Y., is now president of The Electrical Distributors Company, Philadelphia, Pa., which represents electrical manufacturers in the Philadelphia, Pa.; Baltimore, Md.; and Washington, D. C., areas.

OBITUARY ....

Henry Metcalf Hobart (A '94, M '99, F '12) consulting engineer, Schenectady, N. Y., died October 11, 1946. Born November 29, 1868, in Boston, Mass., Mr. Hobart was graduated from Massachusetts Institute of Technology with the degree of bachelor of science in electrical engineering in 1889. That same year Mr. Hobart commenced his career with the Thomson-Houston Electric Company, Lynn, Mass. In 1894 he went to Schenectady, N. Y., with the General Electric Company, where he was assistant to C. P. Steinmetz, who, in recommending Mr. Hobart for AIEE membership, called him "one of the best electrical engineers I ever met." He returned to Lynn in 1895, remaining until in 1896 he joined the British Thomson-Houston Company, London, England. In 1900 he became associated with the Union Electricitats Gesellschaft in Berlin, Germany. From 1903 to 1911 he conducted a private consulting practice in London, afterwards returning as consulting engineer to the General Electric Company. He retired in 1940. Mr. Hobart was a lecturer at the University College, London, from 1908 to 1911 and lectured also at Northampton Institute and Faraday House in London, and he was external examiner for the University of Sheffield for a number of years. He delivered the James Forrest Lecture before the Institution of Civil Engineers in 1915. He was author and coauthor of more than a dozen books, some of which were translated into French and German, and was also author of many technical papers. At least 30 patents were issued in his name. In 1936 he received the Samuel Wylie Miller Memorial Medal of the American Welding Society. Mr. Hobart maintained a life-long interest in AIEE activities. He served as manager for 1922–26, vice-president for 1926–28, and chairman of the Schenectady Section for 1914–15. At various times he was member of the following AIEE committees: executive, Edison Medal, Standards, electric machinery, electric welding, research, and transportation. He represented the AIEE on the American Bureau of Welding, the American Engineering Council, the American Standards Association, and the United States National Committee of the International Electrotechnical Commission. He was a member of the American Welding Society, the American Society of Mechanical Engineers, the American Association for the Advancement of Science, the Institution of Civil Engineers, the Institution of Mechanical Engineers, and the Institution of Electrical Engineers.

Sylvester Bedell Way (A '03, F '38) chairman of the boards of the Wisconsin Electric Power Company, the Milwaukee Electric Railway and Transport Company, the Wisconsin Gas and Electric Company, and the Wisconsin Michigan Power Com-

pany, Milwaukee, Wis., died September 20, 1946. Born August 29, 1874, in Philadelphia, Pa., Mr. Way was graduated from Drexel Institute in 1896. He was awarded the honorary degree of doctor of science by that institution in 1942. After a brief period with the Electric Storage Battery Company, Philadelphia, he became chief electrical engineer and superintendent of what is now the Union Electric Light and Power Company, St. Louis, Mo., in 1898. In 1911 he was appointed assistant general manager of the Milwaukee Electric Railway and Light Company. He became vice-president and general manager in 1914 and president in 1925. He relinquished the position of general manager in 1934 and was elected chairman of the board in 1945. It was under Mr. Way's direction that the company's Lakeside power plant was designed and became the first large central station constructed for the exclusive use of pulverized fuel. In 1917 he was given direct supervision of the company's street railway, and to solve the man power problem during World War I, he designed and patented the center truck construction for two-car articulated trains. Mr. Way was director of the First Wisconsin National Bank, the First Wisconsin Trust Company, the Milwaukee Association, the the Milwaukee Community Fund, the Council of Social Agencies, and the Milwaukee Foundation. He was a member of the Engineers' Society of Milwaukee.

William H. Matthies (M '26, F '29) retired systems engineer for Bell Telephone Laboratories, Inc., New York, N. Y., died October 20, 1946, in Hackensack, N. J. Mr. Matthies was born January 28, 1879, in Hamburg, Germany, and was graduated from Massachusetts Institute of Technology in 1902 with the degree of bachelor of science in electrical engineering. After graduation he joined the Western Electric Company as student engineer and was titled equipment engineer in 1903 and as such began to specialize on circuit design work. At that time he worked on the first dial exchange in the United States. He was appointed chief engineer of the company's Berlin, Germany, branch in 1905 and while in that position engineered and placed in service the first common battery switchboards in use in Germany and Scandinavia. He was attached to the staff of the European chief engineer of the company from 1911 to 1916 at Antwerp, Belgium, specializing in machine switching development. He returned to New York in 1916 and was placed in charge of machine switching circuit development. Under his leadership both the crossbar system of dial exchanges and the earlier panel system were developed. In 1921 he was given charge of all local central office systems development and of private branch exchange development and after 1925 continued with that work for Bell Telephone Laboratories. Mr. Matthies held more than 30 patents now in use in Bell equipment. He retired in 1943.

John Winchell Creasey (A '20, M '29) division plant superintendent, plant department, American Telephone and Telegraph Company, St. Louis, Mo., died recently. Born January 10, 1892, in Rich Hill, Mo., Mr. Creasey received the degree of bachelor of science in electrical engineering from the University of Missouri in 1914. After a year as student apprentice with the General Electric Company, Fort Wayne, Ind., Mr. Creasey was employed briefly as electrician for the Chicago, Burlington, and Quincy Railway, North Kansas City, Mo., and by Peet Brothers Manufacturing Company, Kansas City, Kans. He entered the Bell System in 1916 as equipment attendant for the American Telephone and Telegraph Company, Kansas City, Mo., and the following year was transferred to St. Louis as inspector. In 1919 he was made district plant manager in Kansas City and from 1922 to 1924 he was assigned to the general plant manager's office in New York, N. Y. He returned to Kansas City as district plant superintendent in 1924 and in 1929 was transferred to Dallas, Tex., in the same position. Returning to the office of the general plant superintendent in St. Louis in 1933, he served in the office of the general plant superintendent until in 1936 he was made area plant supervisor. After serving as outside plant engineer in New York from 1937 to 1940, he became division plant superintendent in St. Louis.

James Mowton Saunders Waring (A '06) retired member of the engineering firm of Chase and Waring, New York, N. Y., died October 23, 1946. He was born January 8, 1874, in Baltimore, Md., and was graduated from Johns Hopkins University in 1896. After experience with the United Railways and Electric Company, Baltimore, from 1896 to 1901 and later with the Electric Storage Battery Company, Chicago, Ill., he joined the firm of L. L. Summers and Company, consulting engineers of Chicago, Ill., about 1911. During the first world war he served as a colonel in the Ordnance Department at Nitro, W. Va. He became a partner in the firm of Chase and Waring in 1927. In 1939 Mr. Waring was appointed research director of the political philosophy and social science department at Fordham University, New York. During World War II he was adviser to the War Production Board and the New York State and Maryland Planning Commissions. He was a former president of the Virginia Military Institute Alumni Association and belonged to the Army and Navy Club of Washington, D. C.

Hector Tabossi (A '12, M '19) director and manager, electrical department, General Electric Company of South America, Buenos Aires, Argentina, died June 15, 1946. Born October 21, 1889, in Concepcion del Uruguay, Argentina, Mr. Tabossi was graduated from Ohio State University in 1910 with the degree of mechanical engineer in electrical engineering. In 1912 he received the degree of master of electrical engineering from Harvard University. After a year with the General Electric Company, Lynn, Mass., he became assistant electrical engineer with J. G. White and Company, Ltd., London, England. In 1916 he joined the British Thomson-Houston Company, Ltd., Rugby, England, as commercial engineer, and remained with it until in 1930 he was named commercial engineer for the General Electric Company of South America. He was appointed director and chief engineer in 1935, and director and manager of the electrical department in 1938.

Ferris LeRoy Francisco (A '04, M '16) senior partner, Francisco and Jacobus, Engineers and Architects, New York, N. Y., died October 11, 1946. Mr. Francisco was born October 23, 1879, in Cleveland, Ohio. His earliest engineering experience was gained at electric light, railway, and power house construction work from 1896 to 1902. In 1902 he became electrical engineer for the American Tobacco Company, Louisville, Ky., and in 1903 he was named supervising electrical engineer. He was appointed chief engineer for the company in 1905. In 1912 he organized his own consulting firm with F. R. Jacobus as partner. Mr. Francisco was a member of the American Society of Civil Engineers, The American Society of Mechanical Engineers, and of the Union League Club of New York.

James Harrison (A '03, F '25) retired engineer of the Southwestern Bell Telephone Company, St. Louis, Mo., died June 16, 1946. Mr. Harrison was born September 25, 1874, in St. Louis, and after attending Washington University for a brief period was graduated in 1896 from Harvard University with the degree of bachelor of science in electrical engineering. He joined the Kinloch Telephone Company, St. Louis, in 1897 as timekeeper and surveyor on conduit work and in 1898 was named cable inspector. The following year he was made wire chief and assistant chief engineer, and in 1906 he was appointed chief engineer. He assumed the additional duties of assistant general manager in 1917. When the company was merged with the Southwestern Bell Telephone Company in 1923, he was titled engineer in the general engineering department. He retired from the Bell company in 1933. During the war he was active in the Victory Loan drives.

Clarence Ray Herrington (A '45) resident engineer, Southwestern Public Service Company, Amarillo, Tex., died September 5, 1946. He was born May 10, 1900, in Kansas City, Mo. Mr. Herrington had been employed by the Public Service company since 1917. With previous experience in meter and relay testing, switchboard wiring, and substation and power plant construction and maintenance,

he was made foreman of power plant and substation construction and maintenance in 1939. He became supervisor of substation and power plant maintenance in 1942 and resident engineer at Carlsbad, N. Mex., in 1944. He was a member of the Southwestern Metermen's Association.

Milton P. Kerr (M '41) assistant general manager, Wheeling (W. Va.) Electric Company, died October 7, 1946. He was born January 4, 1889, in Allegheny, Pa. After brief employment as timekeeper and assistant superintendent of the West Penn Electric Company, Pittsburgh, Pa., he was employed as material procurement man and record clerk by the Wheeling Electric Company in 1914. He served in the United States Army in 1918 and returned in 1919 to the Wheeling company as assistant to the general superintendent. He was sucessively construction superintendent and transmission superintendent for the Ohio Power Company, Canton, from 1923 to 1935. He returned to the Wheeling company in 1936 as general superintendent and recently was appointed assistant general manager.

Ranjit Singh Jain (M '39) professor of electrical engineering, Benares Hindu University, Benares, India, died in April 1946. Professor Jain was born October 3, 1892, in Delhi, India, and was graduated from the University of Illinois with the degree of bachelor of science in electrical engineering in 1915. He was assistant engineer with the Reliance Electric and Engineering Company, Cleveland, Ohio, from 1915 to 1919 and in the latter year was appointed manager of engineering for India. He returned to India in 1921 and was appointed professor at Benares Hindu University. He was acting head of the university in 1939. Professor Jain was an Associate of the AIEE from 1916 to 1937 and was a member of the Institution of Electrical Engineers.

Roy Clarence Arter (A '39) chief engineer, North Electric Manufacturing Company, Galion, Ohio, died September 22, 1946. He was born November 13, 1889, in Galion, Ohio, and attended Ohio Northern University. Mr. Arter entered the employ of the North Electric Company in 1911 as wireman, tester, and draftsman. He was named resident factory engineer in 1914 and engineer in charge of design, production, and sales in 1918. In 1922 he was appointed chief engineer.

MEMBERSHIP . .

Recommended for Transfer

The board of examiners, at its meeting of October 17, 1946, recommended the following members for transfer to the grade of membership indicated. Any objection to these transfers should be filed at once with the secretary of the Institute.

To Grade of Fellow

Bushman, A. K., manager, application & serv. engg. div., apparatus dept., General Elec. Co., Schenectady, N. Y.
Concordia, C., analytical engineer, central station engr. divs., General Elec. Co., Schenectady, N. Y.
Edwards, M. A., assistant engineer, general engg. & const. lab., General Elec. Co., Schenectady, N. Y.
Foust, C. M., division engineer, high voltage & nucleonics div., General Elec. Co., Schenectady, N. Y.
Johnson, E. E., manager of engineering, apparatus dept., General Elec. Co., Schenectady, N. Y.
Kilbourne, C. E., designing engineer, motor & generator div., General Elec. Co., Schenectady, N. Y.
Nance, H. H., engineer, long lines dept., Amer. Tel. & Tel. Co., New York, N. Y.
Suits, C. G., vice-pres. & director, General Elec. Research Lab., Schenectady, N. Y.
White, W. C., electronics engineer, research lab., General Elec. Co., Schenectady, N. Y.
9 to grade of Fellow

To Grade of Member

Agner, O. B., sales engineer, Westinghouse Elec. Corp., Philadelphia, Pa.
AtLee, R. Y., engineer, Amer. District Teleg. Co., New York, N. Y.
Baker, H. W., engineer, service dept., Independent Elec. Machy. Co.; owner, designer engineer, Baker Motion Picture Apparatus Co., Kansas City, Mo.
Bostock, G. M., assistant to mgr., switchgear div., General Elec. Co., Philadelphia, Pa.
Bueche, H. S., prof. electrical engineering, Villanova College, Villanova, Pa.
Buss, R. R., assistant prof. elec. engg., Northwestern University, Evanston, Ill.
Campbell, R. E., engineer, Commonwealth Edison Co., Chicago, Ill.
Chernoff, J. L., electrical engineer, U. S. Bureau of Reclamation, Denver, Colo.
Corden, H. F., general office engineer, dept. of power, Tennessee Valley Authority, Chattanooga, Tenn.
Gross, G. J., assistant engineer, Locke Insulator Corp., Baltimore, Md.
Gund, R. A., product engineer, General Elec. Co., Schenectady, N. Y.
Heckendorn, H., electrical engineer, Western Elec. Co., Chicago, Ill.
Howard, L. E., district cable specialist, General Elec. Co., Philadelphia, Pa.
Hunter, A., electrical designer, Electro-Hydraulics (Messier), Ltd., Warrington, Lanc., England.
Kelley, F. B., supt. electrical construction, Kansas City Pr. & Lt. Co., Kansas City, Mo.
Kurz, C. G., senior electrical engineer, U. S. Maritime Comm., Washington, D. C.
Lebenbaum, P., Jr., electrical engineer, General Elec. Co., Lynn, Mass.
Loustaunau, J. J., chief electrical engineer, E. B. Badger & Sons, Co., Boston, Mass.
Lukens, A. F., industrial engineer, General Elec. Co., Lynn, Mass.
Lutz, S. G., chairman, dept. of elec. engg., New York University, New York, N. Y.
McCartney, C. B., consulting engineer, Box 1050, St. Petersburg, Fla.
Miller, N. C., instructor, University of Minnesota, dept. of elec. engg., Minneapolis, Minn.
Miller, R. E., research & development engineer, Curtis Development & Mfg. Co., Milwaukee, Wis.
Olmsted, F. A., electrical engineer, Erik Floor & Associates, Chicago, Ill.
Pierce, R. M., vice-president in charge of engineering, WGAR, WJR, KMPC, Cleveland, Ohio.
Preisel, E. A., electrical engineer, U. S. Navy Yard, New York, N. Y.
Redding, F. A., supervisor, meter section, frequency change dept., Southern Calif. Edison Co., Ltd., Alhambra, Calif.
Reichel, A. N., chief estimator, Busch Bros., Inc., Englewood, N. J.
Richards, K. W., electrical engineer, Dow Chemical Co., Midland, Mich.
Roos, W. B., supt., laboratory, Mexican Lt. & Pr. Co.; prof. elec. engg., Mexican National University, Mexico, D. F., Mex.
Sanger, J. H., electrical engineer, Walter Walkling & Associates, Denver, Colo.
Springer, C. B., assistant engineer, Locke Insulator Corp., Baltimore, Md.
Stefanetti, H. J., assistant supervisor, Pacific Gas & Elec. Co., Marysville, Calif.
Stobbe, J. A., consultant, 30 Broad St., New York, N. Y.
Torian, J. T., member of technical staff, Bell Tel. Laboratories, Inc., New York, N. Y.
VanWambeck, S. H., associate professor, dept. of elec. engg., Washington University, St. Louis, Mo.
Vowels, R. E., State Electricity Comm., Parbury House, Brisbane, Queensland, Australia.
Williams, C. L., chief engineer, Compania Colombiana de Electricidad, Barranquilla, Colombia, S. A.
Williams, E. M., associate prof. elec. engg., Carnegie Institute of Technology, Pittsburgh, Pa.
Yates, J. E., safety supervisor, Pacific Pr. & Lt. Co., Portland, Oreg.
Yavitz, M., engineer, S. C. Sachs Co., Inc., St. Louis, Mo.
41 to grade of Member

Applications for Election

Applications have been received at headquarters from the following candidates for election to membership in the Institute. Any member objecting to the election of any of these candidates should so inform the secretary before December 21, 1946, or February 21, 1947, if the applicant resides outside of the United States or Canada.

To Grade of Member

Abbett, G. W., Abbett Elec. Co., San Francisco, Calif.
Abray, C. H., Canada Coach Lines, Ltd., Hamilton, Ont., Canada.
Adams, F. H., J. S. Clark Equip. & Mfg. Co., El Paso, Tex.
Allen, M., Messrs. Ashanti Goldfields Corp., Ltd., London, England.
Annis, R. B., R. B. Annis Co., Indianapolis, Ind.
Auten, L. D., R.D. #3, Fredericktown, Ohio.
Ayers, O. T., Jr., Florida Pr. & Lt. Co., Miami, Fla.
Bisbee, G. A. W., E. I. duPont de Nemours, Wilmington, Del.
Bostwick, W. G., Precision Welder & Machine Co., Cincinnati, Ohio.
Brewer, N. R., Sinclair Refining Co., East Chicago, Ill.
Byrd, W., Jr., General Cable Corp., Boston, Mass.
Crider, J. M., U. S. Naval Ship Yard, Charleston, S. C.
Crossan, T. E., Virginia Elec. & Pr. Co., Richmond, Va.
Cutler, C. W., General Elec. Co., Chicago, Ill.
Derry, W. H., Aluminum Co. of America, New Kensington, Pa.
Dunn, R. E., Sanderson & Porter, Saltville, Va.
Fitch, L. C., Public Service Elec. Co., Irvington, N. J.
Forbes, H. E., Alabama Pr. Co., Tuscaloosa, Ala.
Garrison, E. W., Jr., Usco Pr. Equipment Corp., Birmingham, Ala.
Geohegan, W. A., Cornell Univ. Medical College, New York, N. Y.
Green, R. E., West Va. Pulp & Paper Co., Charleston, S. C.
Greene, B. F., 225 W. 34th St., New York, N. Y.
Herr, D. L., Control Instrument Co., Inc., Brooklyn, N. Y.
Humphrey, G. W., B. F. Goodrich Chemical Co., Louisville, Ky.
Iberg, E., The Austin Co., Cleveland, Ohio.
Kist, C., Dept. of Water & Pr., Los Angeles, Calif.
Lovell, W. E., Univ. of Saskatchewan, Saskatoon, Sask., Canada.
McIntire, M. M., Bureau of Reclamation, Phoenix, Ariz.
Mercer, W. R., Raytheon Mfg. Co., Waltham, Mass.
Mitchell, J. G. (Re-election), 7720 Sheridan Ave., Chicago, Ill.
Rawlinson, R., Campbell & Isherwood, Ltd., Liverpool, England.
Richart, R. R., McGraw-Hill Publishing Co., New York, N. Y.
St. John, L. E., General Aniline & Film Corp., Binghamton, N. Y.
Schoenhaar, L. H., Consolidated Gas & Elec. Co., Baltimore, Md.
Simpson, A. E., 850 West Hastings St., Vancouver, B. C., Canada.
Spurgeon, S. J., Continental Gin Co., Birmingham, Ala.
Srinivasan, A., Illinois Institute of Technology, Chicago, Ill.
Stocker, O., Kellogg Co., Battle Creek, Mich.
Stoll, A. L., Toledo Edison Co., Toledo, Ohio.
Strickland, H. A., Jr., The Budd Co., Detroit, Mich.
Swanson, D. H., Line Material Co., Milwaukee, Wis.
Taylor, C. E., Delaware Pr. & Lt. Co., Wilmington, Del.
Taylor, P., Central Pr. & Lt. Co., Corpus Christi, Tex.
Thompson, C. F., Sperry Gyroscope Co., Great Neck, N. Y.
Trainor, J. J., Public Service Co. of Ind., Inc., Indianapolis, Ind.
Tyler, R. L., Central Arizona Lt. & Pr. Co., Phoenix, Ariz.
Trier, R. T., Superior Switchboard and Devices Co., Canton, Ohio.
Twining, G. E., Quinton Engineers, Ltd., Los Angeles, Calif.
Ulrich, V. K., Hytron Radio & Electronics Corp., Salem, Mass.
Waldrip, L. H., L. H. Waldrip Co., Cleveland, Ohio.
Warren, C. L., U. S. Army Engineers, Seattle, Wash.
Williams, J. E., Irvington Varnish & Insulator Co. of Canada, Ltd., Hamilton, Ont., Canada.
Williamson, G. P., Crocker-Wheeler Elec. Mfg. Co., Ampere, N. J.
Wright, N. N., Ferranti Elec., Montreal, Que., Can.
Yevick, J. G., Potomac Elec. & Pr. Co., Washington, D. C.
Zavales, C. T., Westinghouse Elec. Corp., Baltimore, Md.
56 to grade of Member

To Grade of Associate

United States and Canada

1. North Eastern

Baschnagel, R. W., Rochester Gas & Elec. Corp., Rochester, N. Y.
Bence, M. W., Int'l General Elec. Co., Inc., Schenectady, N. Y.
Beuerman, R. A., DuPont Co., Buffalo, N. Y.
Bourdon, A. A., Norton Co., Worcester, Mass.
Estes, J. V., Westinghouse Elec. Corp., South Boston, Mass.
Evans, L. W., Naval School (General Line), Newport, R. I.
Gleason, G. F., Rochester Gas & Elec. Corp., Rochester, N. Y.
Hudgings, D. W., III, Stromberg-Carlson Co., Rochester, N. Y.
Johnson, W. G., Westinghouse Elec. Corp., Springfield, Mass.
Joshi, M. G., General Elec. Co., Pittsfield, Mass.
Klarman, K. J., Union College, Schenectady, N. Y.
Mazur, V. N., Westinghouse Elec. Co., Buffalo, N. Y.
McDonald, J. M., Jr., General Elec. Co., Pittsfield, Mass.
Mellichamp, L. R., Jr., General Elec. Co., Bridgeport, Conn.
Miller, A. E. (Re-election), Westinghouse Elec. Corp., New Haven, Conn.
Offensend, L. D., Eastman Kodak Co., Rochester, N. Y.
Rich, M. (Re-election), Chas. T. Main, Inc., Boston, Mass.
Skalnik, J. G., Yale University, New Haven, Conn.
Stallings, H. B., Clark Controller Co., Buffalo, N. Y.
Talbot, C. G., General Elec. Co., Schenectady, N. Y.
Theleen, D. J., General Elec. Co., Schenectady, N. Y.
Wallace, J. T., A. S. Hamilton, Jr., Consulting Engineers, Rochester, N. Y.

2. Middle Eastern

Adams, H. W., Rural Electrification Admn., Washington, D. C.
Barnette, G. L., Hercules Powder Co., Wilmington, Del.
Bogle, E. R., Jr., General Elec. Co., Philadelphia, Pa.
Bovenizer, W. N., Line Material Co., Zanesville, Ohio.
Carlson, M. E., General Elec. Co., Erie, Pa.
Cropley, W. D., The Hoover Co., North Canton, Ohio.
Crouch, W. G., AF, War Dept., Wright Field, Dayton, Ohio.
De Wyer, G. A., Reliance Elec. & Engg. Co., Cleveland, Ohio.
Dobson, D. R., Stackpole Carbon Co., East Pittsburgh, Pa.
Fagge, W. E. (Re-election), P. O. Box 35, Paoli, Pa.
Felton, W. W., The Franklin Institute, Philadelphia, Pa.
Fesler, J. F., Carbide & Carbon Chem. Corp., S. Charleston, W. Va.
Gerrish, D. I., E. I. duPont de Nemours, Wilmington, Del.
Goessel, E., I.T.E. Circuit Breaker Co., Philadelphia, Pa.
Goggins, J. W., Design Service Co., Cleveland, Ohio.
Harr, J. L., Koppers Co., Inc., Pittsburgh, Pa.
Heath, B. W., Reliance Elec. & Engg. Co., Cleveland, Ohio.
Jalen, H. J., Elect. Controller & Mfg. Co., Cleveland, Ohio.
Jokl, A., Westinghouse Elec. Corp., E. Pittsburgh, Pa.
Kelch, R. V., The Ohio Power Co., Canton, Ohio.
Kuhn, M. S., Jr., American Ship Building Co., Cleveland, Ohio.
Landes, L. G., Natl. Advisory Comm. for Aeronautics, Cleveland, Ohio.
Larr, R. B., North Elec. Mfg. Co., Galion, Ohio.
Lucal, C. L., The Ohio Pr. Co., Canton, Ohio.
Malick, F. S., Westinghouse Elec. Corp., Pittsburgh, Pa.
McKelvey, W. O., Koppers Co., Inc., Pittsburgh, Pa.
McMillen, J. R., Ohio Pr. Co., Canton, Ohio.
McMillon, A. B., Jr., General Elec. Co., Erie, Pa.
Miller, W. D., Sperry Gyroscope Co., Inc., Philadelphia, Pa.
Nelson, P. C., Westinghouse Elec. Corp., East Pittsburgh, Pa.
Niemi, W., Leece-Neville, Cleveland, Ohio.
Olin, J. R., General Elec. Co., Erie, Pa.
Parks, G. H., Jr., Westinghouse Elec. Corp., E. Pittsburgh, Pa.
Pascoe, R. J., Westinghouse Elec. Corp., East Pittsburgh, Pa.
Roth, L. M., Carbide & Carbon Chem. Corp., Charleston, W. Va.
Royston, W. W., Air Materiel Command, Wright Field, Dayton, Ohio.
Schurr, C. A., Elec. Controller & Mfg. Co., Cleveland, Ohio.
Seevers, G. E., The Ohio Pr. Co., Canton, Ohio.
Spicer, G. W., I.T.E. Circuit Breaker Co., Philadelphia, Pa.
Sultzbach, R. L., I.T.E. Circuit Breaker Co., Philadelphia, Pa.
Thacker, H. B., Westinghouse Elec. Corp., E. Pittsburgh, Pa.
Thistle, J. R., Leeds and Northrup Co., Phila. Pa.
Tung, I-J., Reliance Elec. & Engg. Co., Cleveland, Ohio.
Westhoefer, D. L., U. S. Naval Ordnance Plant, Canton, Ohio.
Williams, C. M., Ohio Power Co., East Canton, Ohio.
Williams, T. J., The Ohio Power Co., Canton, Ohio.

3. New York City

Armour, B. J., 1201 University Ave., New York, N. Y.
Barton, C. A., U. S. Rubber Co., New York, New York, N. Y.
Berman, R., Communication Measurements Lab., New York, N. Y.
Brown, L. C., Westinghouse Elec. Corp., Bloomfield, N. J.
Fleisher, M., 278 Legion Street, Brooklyn, N. Y.
Giantvalley, J. R., AIEE, New York, N. Y.
Hames, G. J., Cutler-Hammer, Inc., New York, N. Y.
Hauser, R. C., Ebasco Services, Inc., New York, N. Y.
Hayne, C. A., Int'l Standard Elec. Corp., New York, N. Y.
Hill, D. M., Chemical Construction Corp., New York, N. Y.
Lattin, B. C., 3 Bretton Road, Scarsdale, N. Y.
Margolis, H. B., American Gas & Elec. Service Corp., New York, N. Y.
Merer, G. E., Watson Laboratories, Red Bank, N. J.
O'Neill, B. J., General Aviation Equipment Co., Inc., New York, N. Y.
Polak, A. H., Consolidated Edison Co. of N. Y., New York, N. Y.
Rubin, M. (Re-election), 66 Court St., Brooklyn 2, N. Y.
Schiff, N., 858 Faile St., New York, N. Y.
Schwartz, G. J., Arma Corp., Brooklyn, N. Y.
Shapiro, H., Sanderson & Porter, New York, N. Y.
Slezak, S. T. (Re-election), Bell Tel. Lab., Inc., New York, N. Y.
Smelile, L., 6 Burns St., Forest Hills, N. Y.
Trenkamp, A., Jr., 39 Park Ave., Maplewood, N. J.
Watt, G. J., Sperry Gyroscope Co., Garden City, N. Y.

4. Southern

Anest, N., TVA, Chattanooga, Tenn.
Bradley, J. A., Jr., TVA, Knoxville, Tenn.
Ellis, C. J., General Elec. Co., Chattanooga, Tenn.
Hammer, W. A., Jr., TVA, Chattanooga, Tenn.
Hengy, G. C., Geo. C. Hengy & Co., Shreveport, La.
James, W. F., Usco Pr. Equipment Corp., Birmingham, Ala.
Kirkland, J. W., Gulf States Utilities Co., Lake Charles, La.
Lear, W. E., University of Alabama, University, Ala.
Meier, J. A., Florida Elec. Supply, Inc., Tampa, Fla.
Moore, E. V., Sr., Farnsworth Bldg. Co., Memphis, Tenn.
Riall, E. C., Jr., Southwestern Gas & Elec. Co., Shreveport, La.
St. Dizier, H. A., Gulf States Utilities Co., Lake Charles, La.
Taylor, J. D., General Elec. Co., New Orleans, La.
Van Os, J. H., Tulane University, New Orleans, La.
Wright, E. C., Cities Service Refining Corp., Lake Charles, La.
Wright, R. F., Westinghouse Elec. Corp., Chattanooga, Tenn.

5. Great Lakes

Bloom, C. J., Illinois Bell Tel. Co., Chicago, Ill.
Casey, J. E., Leeds & Northrup Co., Chicago, Ill.
Colip, C. D., Colip Bros., Inc., South Bend, Ind.
Cook, L. D. (Re-election), Commonwealth Edison Co., Chicago, Ill.
Curtis, C. L., Iowa State College, Ames, Iowa.
Damidovich, J. F. (Re-election), Western United Gas & Elec. Co., Aurora, Ill.
Fleming, R. T., General Elec. Co., Detroit, Mich.
Goss, W. R., General Elec. Co., Fort Wayne, Ind.
Hari, A., Western United Gas & Elec. Co., Aurora, Ill.
Hickman, G. K., Wisconsin Elec. Pr. Co., Milwaukee, Wis.
Holcomb, W. L., American Tel. & Tel. Co., Chicago, Ill.
Kepper, C. J., Square D Co., Milwaukee, Wis.
Knuth, N. W., Western United Gas & Elec. Co., Aurora, Ill.
Little, M. C., Western United Gas & Elec. Co., Aurora, Ill.
Loye, J. S., Wood Conversion Co., Cloquet, Minn.
McGuire, J. V., Allis-Chalmers Mfg. Co., Milwaukee, Wis.
McKee, J. L., Ill. Bell Tel. Co., Chicago, Ill.
Reed, E. B., Weltronic Co., Detroit, Mich.
Schultz, W. F., Redmond Co., Inc., Owosso, Mich.
Shefveland, A. M., Public Service Co. of No. Ill., Maywood, Ill.
Sigmund, E. F., Allen-Bradley Co., Milwaukee, Wis.
Singh, P., Purdue University, Lafayette, Ind.
Smith, J., Berwind Fuel Co., Superior, Wis.
Srinivasan, S., 1414 East 59 St., Chicago, Ill.
Stout, E. R., 221 East 43rd Ave., Gary, Ind.
Topczewski, E. A., Cutler-Hammer, Inc., Milwaukee, Wis.
Tunell, L. E., Public Service Co. of No. Ill., Joliet, Ill.
Whittaker, E. C., Sangamo Elec. Co., Springfield, Ill.
Wiprud, R. B., Westinghouse Elec. Corp., Duluth, Minn.

6. North Central

Eyer, J. M., Bureau of Reclamation, Denver, Colo.
Larson, N. M., U. S. Bureau of Reclamation, Denver, Colo.
Law, R. K., U. S. Bureau of Reclamation, Denver, Colo.
Lawrence, W. D., Jr., U. S. Bureau of Reclamation, Denver, Colo.
Peterson, H., Peterson Co., Denver, Colo.
Roe, R. A., Continental Oil Co., Casper, Wyo.
Sinclair, K. M., U. S. Bureau of Reclamation, Denver, Colo.

7. South West

Batterson, C. C., Lt., 4413 Rusk Ave., Houston, Tex.
Craig, R. M., Gulf States Utilities Co., Beaumont, Tex.
Duesterhoeft, W. C., Jr., University of Texas, Austin, Tex.
Emfinger, N. W., Texas Elec. Service Co., Ft. Worth, Tex.
Hassinger, C. V., LCRA, Austin, Tex.
Hobusch, W. G., Moloney Elec. Transformer Co., St. Louis, Mo.
Lewis, J. P. (Re-election), Aluminum Co. of America, Kansas City, Mo.
Loper, J. F., LCRA, Austin, Tex.
Slade, W. C., Jr., Southwestern Public Service Co., Denver City, Tex.
Sloan, C. E., LCRA, Austin, Tex.
Smith, H. W., The University of Texas, Austin, Tex.
Steele, E. S. (Re-election), Dept. of Public Works, City Hall, Ft. Worth, Tex.
Voter, E. G., Southwestern Bell Tel. Co., Dallas, Tex.
Wanja, L. F., Texas Pr. & Lt. Co., Dallas, Tex.
Woods, K. H., Southwestern Public Service Co., Plainview, Tex.

8. Pacific

Breuer, J. R., General Elec. Co., San Francisco, Calif
Conrey, D. W. (Re-election), Southern Calif. Tel. Co., Los Angeles, Calif.
Coombs, G. F., California Elec. Works, San Diego, Calif.
Cornell, L. P., Jr., The Pacific Tel. & Tel. Co., San Francisco, Calif.
Dalzell, L. B., Southern Calif. Tel. Co., Los Angeles, Calif.
Devine, G. V., The Pacific Tel. & Tel. Co., San Francisco, Calif.
Dixon, A. A., Modesto Irrigation District, Modesto, Calif.
Dolly, W. W., Supervisor of Ships Office, USN, Terminal Island, Calif.
Emerson, F. E., Railroad Comm. of the State of Calif., San Francisco, Calif.
Hall, R. V., Pacific Gas & Elec. Co., Fresno, Calif.
Lindquist, W. W., Pacific Gas & Elec. Co., Oakland, Calif.
Mahler, F. C., J. E. Redmond Supply Co., Phoenix, Ariz.
McKay, H. B., The Pacific Tel. & Tel. Co., San Francisco, Calif.
McMurtry, D. P., U. S. AAF, Fairfield Army Air Base, Calif.
Raab, W. J., Calif. Prune & Apricot Growers Assoc., San Jose, Calif.
Riha, A. E., Pacific Elec. Mfg. Corp., San Francisco, Calif.
Roadman, C. O., Central Arizona Lt. & Pr. Co., Phoenix, Ariz.
Seaman, R. G., Bechtel Bros. & McCone, Hunting Park, Calif.
Sherwood, H. L., Pacific Gas & Elec. Co., Oakland, Calif.
Sipe, H. T., Pacific Gas & Elec. Co., Oakland, Calif.
Slater, J. E., Stolper Elec. Co., Burbank, Calif.
Stand, T. R., Station KFRC, San Francisco, Calif.
Tarratt, H. P., County of San Mateo, Court House, Redwood City, Calif.
Thomas, R. E., Weihe, Frich & Kruse, San Francisco, Calif.

9. North West

Black, R. C., General Elec. Co., Seattle, Wash.
Dole, R. S., Clearwater Valley Lt. & Pr. Assoc., Lewiston, Idaho.
Loring, A. E., U. S. Government, Fort Douglas, Utah.

10. Canada

Garratt, C., Ferranti Elec. Ltd., Mount Dennis Toronto, Ont., Canada.
Holding, J. D., Canadian Westinghouse Co., Hamilton, Ont., Canada.
Painter, G. W., Canadian General Elec. Co., Ltd., Toronto, Ont., Canada.
Sorby, W. O., Canadian Westinghouse Co., Ltd., Montreal, Que., Canada.

Elsewhere

Chinea, C., R., Nicaro Nickel Co., Nicaro, Oriente, Cuba.
Corbin, G. F., Ketchikan Public Utilities, Ketchikan, Alaska.
Ingram, G. W., Shell Petroleum Co., Ltd., London, England.
Pettit, R. D., The British Thomson-Houston Co., Ltd., Rugby, England.
Toepfer, H. J., Shanghai Pr. Co., Shanghai, China.
Wanger, W., Brown, Boveri & Co., Ltd., Baden, Switzerland.

Total to grade of Associate
United States and Canada, 189
Elsewhere, 6

OF CURRENT INTEREST

Second National Electronics Conference Held in Chicago, October 3-5

With an advance registration of about 1,200 and a final registration of approximately 2,100, the second National Electronics Conference was held October 3–5, 1946, at the Edgewater Beach Hotel in Chicago. During the $2^1/_2$-day conference 19 technical sessions—running as many as 5 in parallel—accommodated 57 lists of technical subjects which were presented in various forms ranging from formal technical papers to extemporaneous reports and panel discussions.

A popular special feature of the conference program was the series of equipment exhibits and demonstrations which filled two large halls in the portion of the building set aside for conference activities. Other special features of the program included two huge general luncheons, one addressed by President Frederick L. Hovde of Purdue University on the subject, "Science, Politics, and the National Welfare"; the other of which was addressed by Vice-President C. G. Suits (M '41) of the General Electric Company on the subject, "Physics of Today Becomes the Engineering of Tomorrow." President Hovde's address will be found on pages 554–6 and that of Doctor Suits will appear in an early issue of *Electrical Engineering*. A joint meeting of the Chicago Sections of the AIEE and the IRE was held Friday evening in collaboration with the National Electronics Conference, at which Doctor J. O. Perrine, assistant vice-president of the Bell Telephone Laboratories, Inc., presented one of his notable animated demonstration lectures "Radar and Microwaves."

1947 CONFERENCE ANNOUNCED

Preliminary plans for the third National Electronics Conference already have been completed and the announcement made that the conference will be held at the Edgewater Beach Hotel in Chicago, November 3–5, 1947. Tentative plans also were announced for the fourth conference, to be held at the same place and at approximately the same date in 1948. It is expected that these future conferences will be of the same nature as the two already held, except that additional attention and probably additional space will be devoted to equipment exhibits and demonstrations. It is the announced plan and objective of the conference sponsors that the program continue to offer an industry-wide opportunity for the interim reports and discussion of any current electronics subject matter regardless of the specialized subdivision of the electrical field, from which the information may become available.

On the basis of present scheduling, the AIEE National Midwest Meeting will be held at the Congress Hotel in Chicago the week of November 3–7, 1947. It is expected that the programs of these two important and somewhat interrelated meetings will be appropriately co-ordinated and correlated, so that AIEE members and others interested in attending one of them will have the opportunity also to take advantage of the extended range of subject matter offered by the other.

ORIGIN OF NEC

The organization of the first National Electronics Conference held in Chicago, October 5–7, 1944, was the outgrowth of the initiative and imagination of a group of ten men who met March 23, 1944, to consider some of the needs and implications growing out of the wartime mushroom growth in the field of applied electronics and related research. The group included P. G. Andres, T. J. Higgins (M '46) J. E. Hobson (M '41), C. S. Roys (M '45), and E. H. Schulz (M '43), all of the Illinois Institute of Technology; R. E. Beam (A '42), A. B. Bronwell (M '44), J. F. Calvert (F '45), all of Northwestern University; Beverly Dudley (M '43) of the McGraw-Hill Publishing Company; and W. O. Swinyard of the Hazeltine Electronics Corporation. Out of the decisions of this group came the first National Electronics Conference, financially underwritten jointly by Northwestern University and Illinois Institute of Technology.

Subsequently, the conference was incorporated under the laws of the state of Illinois as a nonprofit organization: "Serving as a national forum for the presentation of authoritative technical papers on electronic research, development, and application." The officers of National Electronics Conference, Inc., for 1946 are:

President W. O. Swinyard, Hazeltine Electronics Corporation
Executive Vice-President A. B. Bronwell, (M '44) Northwestern University
Vice-President R. E. Beam, (A '42) Northwestern University
Vice-President C. A. Emery, Westinghouse Electric Corporation
Vice-President L. T. Rader (M '43) Illinois Institute of Technology
Treasurer W. M. Ballenger (M '38) General Electric Company
Secretary E. H. Schulz, (M '43) Illinois Institute of Technology

In addition to the foregoing group of officers, the NEC board of directors includes the following:

J. E. Hobson (M '41) Armour Research Foundation
Alfred Crossley Consulting Engineer
R. J. Donaldson (M '44) Commonwealth Edison Company
G. H. Fett (M '38) University of Illinois
R. H. Herrick, Automatic Electric Company
L. S. McPhee, Whiting Corporation
Cullen Moore, Galvin Manufacturing Company
E. O. Neubauer (M '44) Illinois Bell Telephone Company
C. S. Roys (M '45) Illinois Institute of Technology
O. D. Westerberg, Illinois Institute of Technology

The policy and objectives of the National Electronics Conference are reflected further in the following excerpts from the "Foreword" of the "Proceedings" of the 1944 Conference.

"To a large extent, the formation of the National Electronics Conference has been the outcome of the recent rapid growth in the field of applied electronics. Supplementing the important role of electronics in communication, the increased importance of electronic methods and devices in physical measurements, medical application, and the rapid growth of industrial application, have focused public attention on the science and technology of electronics.

"Applications of electronics for military communication and various control operations have been an important factor in promoting the popular appeal of this branch of science which is not so young as commonly is supposed. The origin of electronics—the branch of science and technology which relates to the conduction of electricity in gases or in a vacuum—can be traced back to the observation of the unilateral conductivity of an incandescent lamp with enclosed plate, by Thomas A. Edison in 1882 and of the discovery of the electron in 1895 by J. J. Thomson. Important communication applications of electronic-tube amplifiers had been made by 1915, and by 1923 the radiobroadcasting industry was about to be launched. By 1930, feeble attempts had been made to apply electronic devices in industrial uses but another decade was required to overcome objections to "fragile glass tubes" and to establish a sound footing for industrial electronics.

"While the fundamental principles of electronics and electronic devices are well known and adequately recorded (particularly in the communication field), the diverse applications of this branch of technology too often have been developed in specialized divisions of science. This practice has tended to compartmentalize certain advances, and has made more difficult the dissemination of new discoveries and applications outside of that field in which it originated. It had been evident for some time that a new mechanism, correlating the advances in all branches of electronics, would be desirable."

This is the background of the National Electronics Conference, a "National forum on electronics development and their application," now jointly sponsored by the Illinois Institute of Technology, Northwestern University, The University of Illinois, the Chicago Section of the AIEE,

and the Chicago Section of the Institute of Radio Engineers, with the co-operation of the Chicago Technical Societies Council.

OPENING SESSION

At the opening session, President H. T. Heald of Illinois Institute of Technology delivered the address of welcome which will be published in a future issue of *Electrical Engineering*, as will the speech of Doctor E. U. Condon (M '44) director of the National Bureau of Standards, who spoke on "Electronics and the Future."

Power-Line Telephone Service Soon Available in Six States

Installation has been started on equipment providing rural telephone service over electric power lines in six states, the American Telephone and Telegraph Company, New York, N. Y., recently announced.

Though telephone subscribers in Alabama and Arkansas have been served temperarily by the new method as an experiment, the new service installations will mark the first time that power-line carrier has been utilized to bring telephone service to rural areas beyond the reach of existing telephone lines.

Five Bell telephone companies, two independent telephone companies, four rural power systems financed by the Rural Electrification Administration, and three power companies are parties to the installations. Telephone central offices through which the seven groups of rural subscribers will be served are situated at Manakin, Va.; Nashville, N. C.; Aiken, S. C.; Italy and Lamesa, Tex.; Oak Creek, Colo.; and Cle Elum, Wash.

Equipment for the installations has been manufactured by the Western Electric Company which also has commenced production to meet the expected extension of this type of service.

The telephone company also revealed that an experimental project is scheduled for this fall at Norton Mills, Vt.

Though the projects now in process of installation will be equipped to provide only one telephone channel for each power-line route, power-line carrier apparatus now under development is expected to furnish six speech channels.

Patent Examiners Needed by the United States Patent Office

The United States Patent Office has announced that expansion of its examining corps has created a need for additional patent examiners in grades P-1 and P-2 with starting salaries of $2,644.80 and $3,397.20 respectively.

To qualify for an appointment to a P-1 patent examining position, applicants must have completed a full curriculum of study leading to a bachelor's degree in a college or university of recognized standing including or supplemented by major study in engineering, technology, chemistry or physics; for four years' experience in one of the foregoing fields of such a nature as to demonstrate that the applicants possess an intimate working knowledge of the field involved to the same extent and degree that such a knowledge would have been acquired through college work. To qualify for appointment to a P-2 patent examining position, in addition to the requirements for a P-1 appointment, applicants must show at least one year of responsible experience in the field of chemistry, physics, technology, engineering, or other pertinent work.

After six months of satisfactory service in the P-1 grade, patent examiners are eligible for promotion to grade P-2. Promotions to the higher grades are made from within the examining corps when the employee meets the qualifications for the higher grades.

All interested persons may obtain a Civil Service Commission application form 57, at any first- or second-class Post Office, or Civil Service Commission regional office. Applications should be mailed to the United States Patent Office, Personnel Division, Washington 25, D. C.

Magnetron Adapted to Cooking Purposes

The newly developed "Radarange" which uses ultrahigh frequency electromagnetic power to defrost, heat, or cook a wide variety of foods in a matter of seconds recently was demonstrated by the Raytheon Manufacturing Company, Waltham, Mass.

For use by air lines a special model has been developed which weighs approximately 100 pounds and is designed to defrost and cook a complete meal (which previously has been partially precooked and frozen) in less than a minute. The machine operates on about $4\,^1/_2$ kw from the standard 28-volt d-c power system. The 400-cycle output of motor-generator sets is amplified and rectified to produce direct current at 4,000 volts which is applied to the plate of a continuous-wave magnetron oscillator. The 3,000-megacycle 1,000-watt output of the magnetron is fed to a 4- by 8-inch wave guide by a coaxial cable. The wave guide directs the power to the food which has been placed in a metal tray. The ray acts as a reflector, thereby forcing any stray energy back into the food. Another model for use in drugstores and sandwich shops is ready for production, and a radarange for home use is being developed.

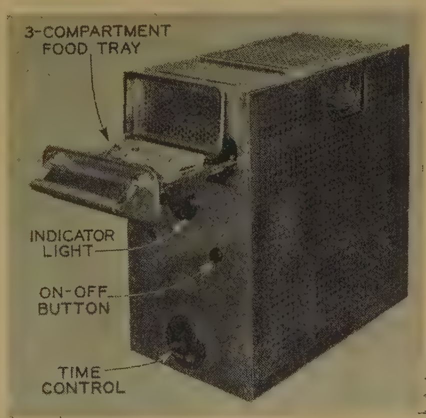


Figure 1. Air line model of the Radarange

This unique cooking device produces hamburgers in 35 seconds. Gingerbread actually can be seen to rise from batter to fluffy muffins in 29 seconds. These muffins have no crusts because the heat is generated in each molecule of the dough and does not have to be conducted from the surface. Hot dogs can be cooked in paper wrappings without burning the paper. This is possible because the water content of the paper is much lower than that of the frankfurter, and water content determines the reaction of each substance to the ultrahigh-frequency energy. As yet no method has been found to boil potatoes, and an egg placed in the range will crack in less than a second. However, it is possible to produce scrambled eggs and baked potatoes.

Portable Instrument Measures Group Opinion

A new instrument designed to measure and indicate the composite opinion of a group of as many as 120 individuals has been announced by the General Electric Company. Called an opinion meter, the instrument enables each person in a group to express automatically, his opinion on any subject. All the individual opinions are summated and registered on a large dial as a single figure in about ten seconds. The instrument has possible use in schools, conference rooms, lecture groups, adult educational organizations, radio stations and political groups.

The opinion meter consists of a large indicating unit and up to 120 hand-held stations. On the individual station is a dial calibrated from 0 to 100, and an adjustable pointer.

The indicating unit is the size of a small suitcase. The cover opens upwards to form a large dial and pointer. Plugs for the string of individual units and for 60-cycle a-c power supply are located on the back of the unit. Inside are electrode and electronic components which perform the measuring and indicating functions.

To register his degree of opinion on a question under discussion, each member of a group turns the pointer on his individual unit to a number on the dial. If strongly in favor of a question, this would be close to 100; if indifferent, 50; and if disapproving, close to 0. The percentage of those in the group who turn their stations to "off," not desiring to express an opinion, also can be determined.

Although the main purpose of the opinion meter is to determine the composite opinion of a group, it also can be used to obtain a ballot. Fifty-fifty opinions are not taken into account in the ballot results, nor are those stations turned to "off."

Exhibitors' Plans Develop for Electrical Exposition

News of exhibits now being developed for the Electrical Engineering Exposition, to be held in New York at 71st Regiment Armory next January 27–31, indicates that the display will open a new channel for information and distribution in a field not hitherto serviced directly by any exposition. The exposition provides an additional feature of interest for those attending the AIEE winter meeting.

For example the introduction of the electronic signal circuit, amplified and relayed to control high voltages, will be featured, and also many new developments in instrumentation specifically applied to the generation and distribution of electric power.

There will be a variety of insulating materials on display, including several classes of plastics, with extensive arrays of products and parts for which they are appropriate, and well-presented information respecting their dielectric properties.

Especially appropriate to this exposition will be the large group of exhibits pertaining to field work, from conductors and pole line equipment of various kinds to transformers, switchgear, and many items of detail—an electronic circuit breaker is one; another is a facsimile demonstration of a lightning arrester at work.

The important needs of maintaining service have not been overlooked, and a number of exhibits will be devoted to maintenance, repair, replacement, and emergency duties, from a newly developed field service truck to a long list of items for in-plant use in time of need.

The armory will be closed to the public, so that a strictly professional and business atmosphere will be maintained. Admission will be by invitation and registration of only qualified visitors.

Medal Presented to Doctor Willis R. Whitney

The fall meeting of the Industrial Research Institute, held October 16–18, at the Westchester Country Club, Rye, N. Y., which brought together more than 100 representatives from the research departments of various industrial organizations, was highlighted by the presentation of the institute's medal to Doctor Willis R. Whitney (A '01) organizer and first director of the General Electric Research Laboratory.

The medal, conferred for the first time since it was established in 1945, was awarded to Doctor Whitney for outstanding contributions to the field of industrial research as a distinguished scientist, a pioneer in the application of organized science to industrial technology, and a beloved and inspiring leader of men. H. W. Graham, senior past president of the Industrial Research Institute, made the presentation at an informal dinner on October 17.

Charles S. Venable, president of the organization, acted as toastmaster at the dinner and introduced the guests at the speakers' table after which Irving Langmuir, associate director of the General Electric Research Laboratory spoke on "Whitney, the Man and Leader." Doctor Langmuir stressed Doctor Whitney's application of serendipity, Webster's definition of which he modified to "the gift of finding *and applying* valuable or agreeable things not sought for," to research. Doctor Whitney, he recalled, always asked those working in the laboratory if they were having fun and never assigned anyone to a project, but let his aides choose according to their own imagination.

In his short address following acceptance of the medal, Doctor Whitney credited Francis Bacon with having started organized research, but on a strictly nationalistic basis with the benefits limited to England alone. Doctor Whitney was emphatic in his opinion that international co-operation among scientists is essential to world peace.

Standards Announced for Fractional-Horsepower Motors

A new program of standardization for fractional-horsepower motors was announced in September by the members of the motor and generator division of the National Electric Manufacturers Association (*EE, Nov '46, p 541*). At that time a basic standard for determining motor rating in co-ordinated terms of horsepower, speed, breakdown torque, and service factor was presented to eliminate certain application difficulties involving rating and performance characteristics.

The standard presented in October provides for uniformity of mounting dimensions and three frame sizes for motors ranging from 1/20 through 3/4 horsepower at 1,800 rpm and including one horsepower at 3,600 rpm. Previously the variety of mounting dimensions and motor sizes necessitated extensive use of many universal mounting bases and adapter plates which resulted in high costs and loss of production time in application of these motors to the products of different manufacturers. The same difficulties have been encountered by service men in attempting to replace a motor of one make with one of another make.

The new dimensions apply to both rigid and resilient or cushion base motors. Those dimensions for which standards have been established include the distance between the axial centerline of the shaft and the bottom of the feet (*D* dimension), distance between the vertical center line and the mounting slot centerlines—end view (*E* dimension), half distance between mounting slot centerlines—side view (*F* dimension), distance from shaft shoulder to nearest mounting hole centerline (*BA* dimension), and diameter of mounting slot or hole (*H* dimension).

In the proposed system for numbering the frames the frame number will be 16 times the shaft height (*D* dimension), and the other dimensions will be constant for any frame size. The three frames will be 42, 56, and 66 with suffix letters used to indicate the same features provided for in the established integral system.

CIGRE Papers Wanted. The Engineering Societies Library has inquired whether any member of AIEE has a set of the papers presented at the 1946 meeting of the International Conference on Large Electric High Voltage Systems, which he would be willing to offer the library. The library has placed an order for a bound volume of the papers, but this will not be published before the end of the year. Correspondence about the papers may be addressed to Ralph H. Phelps, director, Engineering Societies Library, 33 West 39th Street, New York 18, N. Y.

Television at Convention. Television solved the problem of accommodating the overflow audience at the sessions of the 24th annual convention of the National Broadcasters Association held at the Palmer House in Chicago, Ill., recently. Three RCA Image Orthicon television viewing equipments were used to pick up and transmit events in the meeting rooms to 20 RCA Victor television receivers installed in the hotel's exhibition hall for the overflow audience. According to Henry Rhea, manager of RCA television equipment sales, use of the supersensitive Image Orthicon camera permitted pickup with ordinary room lighting, whereas the discomforts, cost, and inconvenience of the special brilliant lighting otherwise required would make such a service impractical.

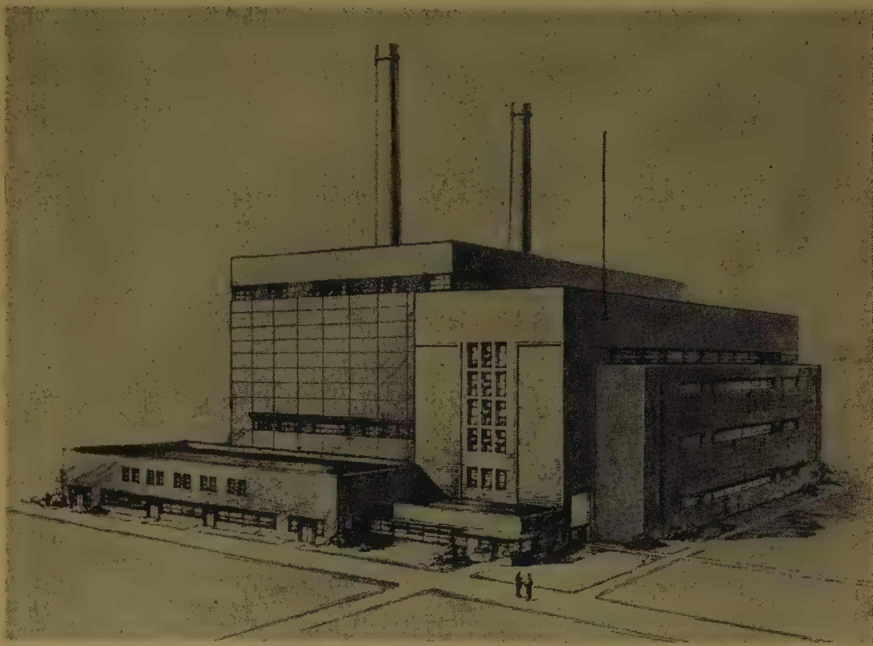
Commercial Air Line Installs Radar. First radar installations in a $300,000 experimental program recently were made by American Airlines, Inc. The flagship *St. Joseph*, a four-engine DC-4 cargo airplane, became the first commercial airplane to be equipped with radar to facilitate instrument landings in bad weather. A $15,000 radar beacon station is being installed on the air line's hangar in St. Joseph, Mo., and a twin-engine DC-3 laboratory ship has been made available for pilot instruction.

Experts Aid Navy Science. Appointment of ten experts to the newly established Civilian Research Advisory Committee has been announced by Secretary of the Navy James Forrestal. Named to the committee which was authorized by the last session of Congress to assist in spurring Naval scientific research are: Richard J. Dearborn (A '16) president, Texaco Development Corporation, New York, N. Y.; Doctor Karl T. Compton (F '31) president of the Massachusetts Institute of Technology, Cambridge; Lewis L. Strauss, United States Naval Reserve, of Kuhn, Loeb and Company, New York, N. Y.; Luis de Florez, USNR, vice-president in charge of engineering, Doubleday and Com-

pany, New York, N. Y.; Doctor Warren Weaver, director of the division of natural sciences, Rockefeller Foundation, New York, N. Y.; Doctor Philip M. Morse, professor of physics, Massachusetts Institute of Technology, who will become director of the Northeastern Regional Laboratory for Atomic Energy Research, Camp Upton, N. Y.; Doctor L. A. DuBridge, president of the California Institute of Technology, Pasadena; Doctor Arthur H. Compton, chancellor of Washington University, St. Louis; Doctor William Sharp McCann, director of the Institute of Medicine, Rochester (N. Y.) University; Doctor Detlev W. Bronk, head of the National Research Council.

Icaroscope Described to Optical Society. The secret of another war-born defensive weapon, the Icaroscope, was revealed by Professor Brian O'Brien (M '28) director of the Institute of Optics of the University of Rochester, N. Y., at the 34th annual meeting of the Optical Society of America held recently in New York. Shaped like a stubby telescope, the Icaroscope enables pilots to find enemy airplanes diving toward them directly in line with the sun. The image of the airplane is formed on a transparent screen made from a phosphor chosen for short afterglow with saturation, while the screen is hidden from the eye. A double rotating screen then closes off the outside light and the pilot sees the image 1/100 second later. By that time the image on the screen is only 20 to 50 times the brilliance of the surrounding sky, as compared with the sun's real brightness of 10,000 to 100,000 that of the surrounding sky, and the airplane is silhouetted on the screen against either the sun or the surrounding sky. The disks are rotated by an electric motor at about 100 cycles per second, so that the images appear with the rapidity of a motion picture. The Icaroscope was developed under contract with the Office of Research and Development as an aid in defense against aircraft attacking from the general direction of the sun. It also was used in observing and photographing the Bikini atom bomb tests.

Generating Station Designed for Rochester



This new $15,000,000 steam electric generating station will be situated on the shores of Lake Ontario at Rochester, N. Y., the Rochester Gas and Electric Corporation has announced. The station, which will have an ultimate capacity of 160,000 kw will be constructed in four units of 40,000 kw each, the first of which is expected to be completed in 1948 at a cost of $8,000,000. Each unit will include one boiler and one turbogenerator with all the necessary auxiliary equipment. The turbine will be operated by steam at 1,250 pounds pressure at the turbine throttle. Station dimensions will be 280 by 250 feet by 100 feet high

OTHER SOCIETIES

Building Conference Organizes Building Officials Foundation

Establishment of the Building Officials Foundation, a co-ordinating and research body, which will work for uniform and modern building codes throughout the United States, was announced at the 31st annual meeting of the Building Officials Conference of America, Inc., held recently in Memphis, Tenn.

The principal functions of the Foundation were given as:

1. To encourage the adoption of the "Basic Building Code" promulgated by the Building Officials Conference of America, Inc., by all communities, to the end that a country-wide uniform code be established and to encourage communities to keep their local codes up to date.

2. To assist communities in the administration of their building laws and regulations.

3. To reconcile local building codes with advances in building to the end that structural design, equipment, construction materials, and methods may be progressively improved in the public interest.

4. To provide uniform testing procedure in consultation with industry in determing the adequacy of testing equipment, and for the evaluation of the integrity of equipment, materials, and methods of construction.

5. To make known the results of tests promptly and in a manner compatible with the needs for such data in administering building rules and regulations.

6. To keep building and other municipal officials concerned with public health and safety, manufacturers, producers, architects, engineers, general and special trades contractors and other builders, real estate dealers and administrators, property managers and appraisers, mortage and equity financiers, government officials concerned with construction, building labor leaders, materials dealers, suppliers and distributors and others in or related to the building industry informed of advances in building requirements and code administration and to supply them with news, informative articles, and other data pertaining to better building and construction practices.

The services of the Building Officials Foundation are designed to function primarily for building officials as administrative agents of building laws and regulations, and for the betterment of the building industry as a whole. The various service instruments are as follows:

1. The Basic Building Code, with its Construction Code and Small Municipality Building Code.

2. The structural bureau to analyze plans and projects of unusual magnitude or design submitted by understaffed municipalities.

3. The testing procedure for the introduction of new devices, materials, and methods meeting the functional requirements of building as provided in the Basic Building Code.

4. The *Official Bulletin* of the Building Officials Foundation containing test reports and recommendations for the use of approved materials, equipment, and construction methods.

5. A monthly illustrated magazine for general distribution to building officials and other interested persons in the building industry.

The foundation is a nonprofit organization and was set up by a voluntary endowment fund contributed by the construction industry. For future expenses the foundation will depend on annual dues of participating members and fees collected from

Future Meetings of Other Societies

American Chemical Society. 111th national meeting, April 14–18, 1947, Atlantic City, N. J. 50th annual meeting, June 16–20, 1947, Atlantic City, N. J.

American Society for Testing Materials. Spring meeting and committee week, February 24–28, 1947, Philadelphia, Pa.

American Society of Mechanical Engineers. Annual meeting, December 2–6, 1946, New York, N. Y.

American Society for X-ray and Electron Diffraction. Winter meeting, December 5–7, 1946, Pittsburgh, Pa.

Institute of Radio Engineers. Annual meeting, March 3–7, 1947, New York, N. Y.

National Exposition of Power and Mechanical Engineering. December 2–7, 1946, New York, N. Y.

tests and the promulgation of standards. The grades of membership will be:

1. Ordinary members will include all building officials.
2. Members of the Building Officials Conference of America.
3. Founder members are subscribers to the voluntary endowment fund.
4. Participating members may be manufacturers, producers, financial institutions, insurance companies, general contractors, and others engaged in the building industry.

IES Lighting Handbook Postponed. Issuance of the Illuminating Engineering Society "Lighting Handbook" originally scheduled for 1946 has been postponed to October 1947. In making the announcement, C. A. Atherton, chairman of the handbook committee, explained that reconversion problems have otherwise occupied members working on the handbook, so that production necessarily must be delayed. Postponement of publication is expected to aid manufacturers in supplying the most recent material for the reference data section. When completed, the 500-page handbook will cover every phase of lighting from the pure physics of light to specific lighting recommendations for stores, offices, homes, factories, and even for television studios and juke boxes. The latest information on light sources, as well as the measurement and control of light will be included.

Civil Engineers Nominate President. Edgar M. Hastings of Richmond, Va., chief engineer of the Richmond, Fredericksburg, and Potomac Railroad Company, has been nominated as the 1947 president of the American Society of Civil Engineers. A graduate of Baltimore City College and Baltimore Polytechnic Institute, Mr. Hastings' entire professional career has been in the field of railroad engineering, and he has been associated with the Richmond, Fredericksburg and Potomac Railroad since 1903. He has been chief engineer since 1922. Mr. Hastings is 64 years old.

Relations With Government Considered by Canadian Council

Collective bargaining, the type of technical employment service provided by the government, and school curriculums made up the agenda of the fourth regular meeting of the Canadian Council of Professional Engineers held in Ottawa, Ontario, in September.

The meeting was concerned with the rate at which experienced staff is leaving the service of the Wartime Bureau of Technical Personnel, and with the attempt of the Executive and Professional Division of the National Employment Service to deal with technical persons, with completely unsatisfactory results. It was agreed to direct the attention of the Minister of Labour to these developments, and at the same time to advise the co-operating organizations with respect to the danger of serious deterioration in the quality of national employment service available to their members.

Recent decisions of the National Labour Relations Board on certification of bargaining units of professional people were reviewed, and it was agreed to discuss certain aspects of these decisions on an informal basis with senior administrative officers of the Dominion Department of Labour. The attention of co-operating organizations was directed to the fact that the Minister of Labour had invited representatives of all provinces to meet in Ottawa in an endeavor to draft a national labor code to serve as a basis for provincial labor legislation.

It was announced that information had been assembled regarding all undergraduate and postgraduate scholarships, bursaries, and the like in the field of science at Canadian universities and that the material was being classified for presentation in a final form for the next meeting. A draft of a brief statement outlining the existing collective bargaining situation in Canada as it affects technical persons and professional engineering and scientific organizations was presented to the meeting, and it was decided to transmit copies to all co-operating organizations.

U. S.–Canadian RMA Meeting. United States and Canadian directors of the Radio Manufacturers Association exchanged information on production, labor relations, and school sound equipment at a joint conference held in Quebec, Quebec, in October. A similar joint conference is planned for next spring in a United States city. During the meeting RMA directors appropriated $10,000 to be spent in the promotion of National Radio Week, November 24–30 in conjunction with the National Association of Broadcasters, as the first step toward a closer liason between the two associations. The RMA also voted to support the manufacturer–agent disposal system being used by the War Assets Administration for surplus electronic equipment and to reject a Department of Commerce proposal that the RMA screen captured German technical documents on electronic developments.

AIME Saunders Medal Awarded. The award of the William Lawrence Saunders Gold Medal for 1947 to LeRoy Salsich, president of the Oliver Iron Mining Company, Duluth, Minn., has been announced by the American Institute of Mining and Metallurgical Engineers. The medal was awarded Mr. Salsich for "his conspicuous success in developing men and methods for the mining and transportation of iron ore; for his significant contribution, as operating head of the world's largest iron mining enterprise, to the nation's production of steel so vital to victory in World War II." Mr. Salsich has been associated with the Oliver company, a subsidiary of the United States Steel Corporation, since his graduation from the University of Wisconsin in 1901. He was born in 1879 in Hartland, Wis.

EDUCATION . . .

Cornell Receives Navy Gift. An outright gift of a laboratory unit of Diesel engines and related equipment valued at more than $2,000,000 recently was made by the Navy Department to Cornell University, Ithaca, N. Y. The equipment, accumulated during the war in the course of training almost 3,000 Naval officers, is described as a completely integrated installation adapted to professional instruction and research in the theory, operation, and maintenance of Diesel engines. The main laboratory's 23 engines, the largest weighing 10 tons, range from 1 to 16 cylinders, generating from 15 to 1,700 brake horsepower and represent about 15 different manufacturers. In addition the gift includes the equipment of two auxiliary laboratories, one in which preliminary training in internal combustion was given during the war, and one containing the modern electric equipment used on Naval vessels, as well as air compressors, pumps, refrigeration units, and a communication system. In accepting the gift, Vice-President S. C. Hollister, dean of the college of engineering, said that the equipment not only could be used in existing courses in mechanical and electrical engineering but also would enable the university to organize new elective courses, both graduate and undergraduate.

Georgia School Offers Doctor's Degree. The first engineering and scientific courses in the State of Georgia leading to the degree of doctor of philosophy are being offered by the Georgia School of Technology, Atlanta. The graduate division is scheduling 357 courses which can be pursued by graduate students toward master of science and doctorate degrees. Graduate students applying for the doctor's degree are encouraged to undertake research which will benefit the South from an engineering and industrial viewpoint. The School of Technology also has available financial aids ranging from $600 to $1,800 per academic year to

assist students who would be unable to continue advanced studies because of financial need. These include assistantships, instructorships, and fellowships.

Arkansas Gets Engineering Courses. Four-year courses in electrical, mechanical, and civil engineering have been instituted at Arkansas State College, Jonesboro, in response to the current demand for engineering curricula. Applications are being accepted for enrollment in the quarter beginning November 29, 1946, and for the fall semester of 1947. The campus has adequate living accommodations for single men on a barrack basis. New laboratory equipment is being collected and installed for the engineering department. Most of the college buildings have been constructed since 1932, though the college itself was founded in 1909.

INDUSTRY

Distribution Department Uses Two-Way Radio Communication

The electric distribution department of the Public Service Electric and Gas Company, Newark, N. J., has completed the construction of a two-way radio communication system by means of which supervisors of the electric operating divisions can communicate directly with line crews, trouble men, and field supervisors wherever they may be working.

Seven fixed stations at different cities in New Jersey serve as centers from which orders are transmitted to field crews in the surrounding territories. Each station has 50-watt frequency-modulated transmitters and antennas more than 100 feet high. Receiving equipment also is installed in the fixed stations.

Mobile radio equipment is being installed in more than 150 cars and trucks. These units are equipped with frequency-modulated receivers and 35-watt transmitters to permit crews to report back. Both receivers and transmitters are energized from the storage battery of the car or truck.

A conventional telephone handset with a "push to talk" button is used at both mobile and fixed stations. As all the equipment is designed for the same frequency, it is possible for any mobile unit to contact another mobile unit.

Utility Council Formed. Formation of the New York State Utility Council was the outcome of a recent two-day meeting of nearly 100 executives of gas and electric companies at Saranac Lake, N. Y. The new council, it was decided, will meet informally from time to time to discuss common problems. Robert E. Ginna, vice-president of the Rochester (N. Y.) Gas and Electric Corporation was designated chairman for the coming year. Among the addresses at the recent meeting was one on "Economic Research as a Basis for Selective Load Building" by Constantine Bary (M '43) rate research engineer, for the Philadelphia (Pa.) Electric Company.

Tube Testing Chamber Converted From War Uses

A tube testing chamber originally designed to test tubes for wartime aircraft equipment now has been equipped for multipurpose testing of all types of tubes, the General Electric Company has disclosed.

Capable of simulating a wide range of temperature and altitude conditions within brief periods of time, the over-all equipment for the chamber occupies a space of approximately seven feet by eight feet by 15 feet. Altitude is attained at the rate of 3,000 feet a minute while 100 degrees below zero can be reached in about two hours and 175 degrees above zero in about 45 minutes. Two dials outside the door automatically control, operate, and record, on a 24-hour chart conditions within the chamber.

Leads into and out of the chamber furnish detailed information on tube operation under varied conditions. There are 20 standard leads, 12 pair of thermocouple leads, and three high-frequency coaxial leads connected to a terminal board on one side of the chamber where measurements can be made, it was explained. Four pipes through the walls of the "tube icebox" also make provision for the testing of forced-air and water-cooled tube types.

Amplidyne Booster Inverter Developed for Railway Trains

Announcement of a new amplidyne booster inverter which will provide railway trains with an economical supply of 60-cycle a-c electric power ample for radios, movies, public-address systems, fluorescent lighting, and improved air conditioning has been made by the General Electric Company.

Mounted under the railway car, it changes the car's axle-driven generator or battery from direct to alternating current and gives constant voltage and frequency without excessive losses. The inverter is rated 6.25 kva, 5 kw, and 0.8 power factor to invert direct current of from 90 to 160 volts to 230-volt 3-phase 60-cycle current with voltage and frequency closely held by means of a separately mounted regulator. The amplidyne booster inverter consists of an inverted converter (synchronous converter running from the d-c side) with an amplidyne mounted on the same shaft, operating at approximately 1,800 rpm.

Aluminum Substitutes for Copper. The United States Rubber Company, New York, N. Y., recently announced that it will substitute aluminum for copper in some of its building wire and cable now going into production. This action was taken as a means of alleviating the wire shortage resulting from the temporary scarcity of copper, the company stated, and expressed the belief that it was removing an obstacle to building construction. The aluminum insulated wire will be made in all sizes, as approved by the Underwriters' Laboratories, with the same over-all diameter as equivalent copper sizes.

Sarnoff Honored for 40 Years in Radio. David Sarnoff (M '23) president of the Radio Corporation of America, ventured some predictions on future scientific advances at a dinner given recently in the Waldorf Astoria Hotel, New York, N. Y., to commemorate his 40 years of service to radio. Some of the advances foreseen by Doctor Sarnoff, who admitted that forecasting the developments of science is an almost impossible task, were control of the weather by man, delivery of mail by radio, portable communication sets that will enable one person to talk with another anywhere, transformation of climates by diversion of ocean currents, world-wide television, and use of atomic energy to combat disease. Officials of RCA presented to Doctor Sarnoff the wireless key he used as an operator on April 14, 1912, the day the Titanic sank. On duty at that time at John Wanamaker's, he remained at his key for three days, reporting the names of the 706 survivors as they came in. Other speakers at the dinner included Owen D. Young, retired chairman of the board of the General Electric Company, and Doctor Karl T. Compton (F '31) president of Massachusetts Institute of Technology.

JOINT ACTIVITIES

Engineering Foundation Elects Officers for 1946

Officers for 1946–47 were elected at the annual meeting of the Engineering Foundation with Doctor A. B. Kinzel, vice-president of Union Carbide and Carbon Research Laboratories, Inc., and of the Electro Metallurgical Company, re-elected president of the Foundation. Other officers elected were:

Vice-Chairman—Doctor L. W. Chubb (F '21) director of the Westinghouse Research Laboratories, East Pittsburgh, Pa.

Director—Doctor Edwin H. Colpitts (F '12) retired vice-president of Bell Telephone Laboratories, Inc., New York, N. Y.

Secretary—John H. R. Arms.

Doctor Kinzel will represent the Engineering Foundation on the executive board of the National Research Council.

The Research Procedure Committee will be headed by Doctor Chubb for the coming year. Other members of the committee will be B. A. Bakhmeteff, consulting engineer and professor of civil engineering, Columbia University, New York, N. Y.; J. F. D. Smith, acting president of Purdue University and dean of the engineering

school, Lafayette, Ind.; and F. F. Colcord, vice-president and manager of metal sales for the United States Smelting, Refining, and Mining Company, New York, N. Y.

The executive committee for the year will consist of Doctor Kinzel, J. Schuyler Casey, R. H. Chambers, C. R. Jones (M '30), and O. E. Buckley (F '29). Mr. Arms is secretary of the committee.

The Engineering Foundation is one of the departments of the United Engineering Trustees, Inc., a corporation which was set up jointly by the four national engineering Founder Societies which have an aggregate membership of more than 88,000. The foundation was set up for "the furtherance of research in science and engineering, and the advancement in any other manner of the profession of engineering and the good of mankind." The four Founder Societies are the AIEE, the American Society of Civil Engineers, the American Institute of Mining and Metallurgical Engineers, and the American Society of Mechanical Engineers.

New ASA Director of Information. Frank MacMillen, until recently a member of the staff of the New York *Times* has been appointed director of information for the American Standards Association. He will supervise an expanded educational program designed to inform the public, as well as interested industrial and technical groups, concerning the activities of the association. Mr. MacMillen previously had been on the staffs of the Associated Press, the *Wall Street Journal*, and *Newsweek* magazine.

UET Elects 1946–47 Officers. J. P. H. Perry, vice-president of the Turner Construction Company, New York, N. Y., has been re-elected president of the United Engineering Trustees. Other officers elected are:

Vice-Presidents—Ralph M. Roosevelt, New Canaan, Conn.
William H. Harrison (F '31) vice-president of the American Telephone and Telegraph Company, New York, N. Y.

Treasurer—Albert Roberts of Minerals Separation North American Corporation, New York.

Assistant Treasurer—C. R. Jones (M '30) Eastern transportation manager, Westinghouse Electric Corporation, New York, N. Y.

Secretary—John H. R. Arms.

The United Engineering Trustees is a corporation set up jointly by the four Founder Societies for the advancement of the engineering arts and sciences in all their branches. It functions through two departments: the Engineering Foundation and the Engineering Societies Library.

LETTERS TO THE EDITOR

INSTITUTE members and subscribers are invited to contribute to these columns expressions of opinion dealing with published articles, technical papers, or other subjects of general professional interest. While endeavoring to publish as many letters as possible, Electrical Engineering reserves the right to publish them in whole or in part or to reject them entirely. Statements in letters are expressly understood to be made by the writers. Publication here in no wise constitutes endorsement or recognition by the AIEE. All letters submitted for publication should be typewritten, double-spaced, not carbon copies. Any illustrations should be submitted in duplicate, one copy an inked drawing without lettering, the other lettered. Captions should be supplied for all illustrations.

An Analysis of Electromagnetic Forces

To the Editor:

Reference is made to a letter from Alfred Gronner which appeared in the June 1946 issue of *Electrical Engineering*, pages 300–02, commenting on my article "An Analysis of Electromagnetic Forces" which was published in the October 1945 issue, pages 351–6. In his letter Mr. Gronner, borrowing from certain relativity concepts, offers a mathematical study in support of the popular belief that "parallel currents attract."

While this belief is not an inherent part of conventional electromagnetic theory, it is a natural result of the prevailing methods of interpreting the subject. Following this belief to its logical conclusion leads to the inconsistencies I spoke of in my article, and the combination of relativity and orthodoxy proposed in Mr. Gronner's letter does not resolve them. In fact, it appears to add to their complexity by presenting a derivation for the force between current streams which involves the velocity of the observer. This would make the activities of nature dependent upon the viewpoints of observers, a result opposed to all sense of logic, and one which I was able to eliminate in my analysis. Otherwise one is led to inquire how nature will act in the absence of observers. A situation such as that proposed by Mr. Gronner would shake the very foundations of electrostatics, and leave no basis for electrodynamics except in the mind of the observer. If this conclusion seems a little strong, it is inescapable if one accepts at its face value Mr. Gronner's statement: "F_{2y} is a force *due to the relative motion of the observer and the charges.*" The italics are his, presumably intended to emphasize the conclusion just stated, following which he derives from the value of F_{2y} the formula for the electromagnetic force between current streams. I prefer to believe that electromagnetic forces are natural forces, having a real existence, and not merely existing in the mind of the observer. Nature must be presumed to proceed in an orderly fashion without the requirement of human intervention.

The approach to this subject as outlined in my article is receiving considerable attention by a number of investigators, particularly in the more scientific circles, while there is increasing evidence that even in engineering circles thoughtful individuals are becoming doubtful of the basis for certain prevailing beliefs. F. W. Warburton of the University of Kentucky, in letters to the editor of the *Physical Review*,[1,2] offers formulas for the forces between charges based solely on the relative velocity and acceleration of the charges. I might interpolate here that my article in the October 1945 *Electrical Engineering* was based on, and limited to, the concept of relative velocities, and showed that this concept led to the demonstrable electromagnetic forces; but that I have applied the method to relative acceleration also, with the result that these lead to an explanation of the effects of electromagnetic induction, such as induced voltages, self- and mutual-induction, radiation, and the like.

As to the conventional treatment of Maxwell's concept of the electromagnetic field, definite dissatisfaction is voiced by R. B. Lindsay and H. Margenau, of Brown University and Yale University respectively, who in a book[3] on physics state: "It failed, of course, to meet the demands of the flood of electrical phenomena discovered during the last few decades of the century, but in its very failure it has influenced profoundly the course of present-day physics." In view of such opinions I believe one is justified in questioning the validity of the conventional approach, even when employed by so eminent an authority as J. A. Stratton of the Massachusetts Institute of Technology (my own alma mater) who in his book[4] on the subject (one of the textbooks in the McGraw-Hill international series in physics) starts out in the first paragraph with an acceptance of Maxwell's equations and postulates them as the basis of his development of the subject.

Doubt is simmering in Great Britain also. Alfred O'Rahilly[5] of University College, Cork, has devoted the major portion of a large and masterful volume to this very subject. The terms "doubt" and "simmering" are, perhaps, a bit mild to use in referring to O'Rahilly's strongly stated opinions, but may seem better fitted to some of the comments appearing in the technical press. Reference is made, for instance, to articles by G. W. O. Howe[6,7] in the *Wireless Engineer*, London, and I. A. Robertson[8] in the *Philosophical Magazine*.

There is a substantial but rather scattered literature on the subject, the foregoing references illustrating some of the more recent only. Even such a limited group as this, however, demonstrates the chaotic situation existing with regard to the fundamental theory behind the matter. For instance, references 6 and 8 discuss the same point, which concerns an apparent paradox in the case of charges moving at right angles to each other. Both commentators point out that conventional analysis leads to the conclusion that there is a force on one particle due to the other, but no force on the second due to the first, a result at variance

with Newton's third law of motion. Robertson attempts to explain away this paradox by drawing on the work of Ampere, and arrives at the conclusion that there is no force on either charge. Howe arrives at the same conclusion by an involved and unconvincing application of Maxwell's displacement current concept. The relative velocity concept followed in my article leads to the conclusion that there are equal forces on both charges. It is interesting to note in this connection that my formula (equation 4 in my article) is substantially that of Ampere, except that I use relative velocities and Ampere used absolute velocities, whatever they may mean today.

Howe's other article (reference 7) and Mr. Gronner's comments illustrate another divergence of opinion. In this case Howe attempts to explain the pinch effect as a purely electrostatic phenomenon, while according to Mr. Gronner it is an electromagnetic phenomenon by reason of the motion of the negative charges relative to the observer. But in my article I show that the relative velocity concept explains it as an electromagnetic phenomenon due to the motion of the negative charges relative to the positive charges.

These instances illustrate only some of the chaotic conditions and the shades of opinion existing in presumably well-informed circles. Something should be done to clear up the situation, and that is what I am trying to do, if only to the extent of stirring up the subject.

Echoes of these more technical aspects of the subject have appeared to some extent in the pages of *Electrical Engineering* other than in my article. In the October 1945 issue, page 381, the same issue in which my article appeared, B. Litman of the Westinghouse Electric Corporation, in a letter to the editor, raises a question on the very point at issue between Mr. Gronner and myself. I have often wondered whether Mr. Litman read the article and was satisfied with my explanation. Also in *Electrical Engineering* there have appeared, from time to time, essays for recreation. These have been interesting but many of them would have had no point if there were not certain widely held misconceptions upon which they were able to build. Other articles have appeared in the past, notably those of many years ago by Carl Hering, which I read at the time with much interest and which I am glad to acknowledge had much to do in arousing my interest in the subject.

In the analysis of electromagnetic forces as developed in my article, I arrive at a concept for the effects of relative velocity, which I embody in two propositions. It is presumably these two propositions to which Mr. Gronner refers when he accuses me of postulating new laws of nature. Actually the concept is not new at all, having been formulated by Wilhelm Weber almost exactly 100 years ago. It was developed further by a number of the leading scientists of the latter half of the 19th century, including Clausius and Riemann, and even to some extent by H. A. Lorentz, until obscured by the ideas of Maxwell. Now the wheel turns again. The Maxwellian approach is found wanting, and the older concept is receiving renewed attention. My two propositions, which I developed strictly in accordance with this concept, lead directly to my formula for the force between moving charges. This is the formula mentioned above which, except for using relative velocities instead of absolute velocities, is equivalent to the one Ampere developed as far back as 1823. It proves to be a powerful method, and with our greater knowledge of the nature of matter and electricity, it is hoped that we can forge it into a more complete theory of electromagnetism than did our scientific predecessors.

To continue with a specific consideration of Mr. Gronner's comments, it seems superfluous to point out that the action of current flow must produce forces of a definite and predictable magnitude. Take the case of short-circuit currents, which can produce destructive effects if precautions are not taken to provide adequate support for the conductors carrying these currents. Certainly the dangerous effects possible in such cases cannot be dependent upon the viewpoint of an observer. But application of Mr. Gronner's method and assignment of various velocities to his observer certainly will lead to a wide variety of inconsistent results, as will be obvious. It does not seem necessary to give specific numerical examples.

The trouble with Mr. Gronner's analysis is that he has considered only a special case (that of equal parallel velocities) and attempts to build up a general theory of electromagnetism from it. It is suggested that his method be tried for the general case, in which the velocities of the charges may have any assignable individual values. It then will be found that the analysis cannot be carried out without eliminating the observer and considering the velocity of the charges relative to each other. This case will bring out also the main mathematical error in Mr. Gronner's analysis, which occurs in his equation 18 wherein v^2 is equated to $v_1 \times v_2$. This error is obscured in Mr. Gronner's work, as he does not use the subscripts 1 and 2, but he definitely uses these two velocities as the individual velocities of the two current streams in equations 20 and 21. In his special case they have equal magnitudes but are none the less separate mathematical entities and cannot be derived from v^2. This will be clear in the general case wherein v_1 does not equal v_2, and, therefore, $v_1 \times v_2$ cannot be equal to v^2. In this case there is no means of making the transition from v^2 to $v_1 \times v_2$, and I should like to see Mr. Gronner's method of accomplishing this step.

I might mention, for the benefit of anyone interested in analyzing Mr. Gronner's letter, that there are a number of typographical errors which make the mathematics somewhat difficult to follow. These appear to be truly typographical in nature, and no criticism can be made of Mr. Gronner for their occurrence. On the other hand, I do not believe he has added to the clarity of his case by including the appendix for the derivation of the integral of equation 16. The integrand in equation 16 is a standard form of a quadratic to the minus 3/2 power and is given in any short table of integrals. It yields the result obtained if one makes the approximation that $x/\sqrt{x^2+\beta^2a^2}=1$, when $x=\infty$. This is just as accurate an approximation as $\tanh\ \alpha=1$, when $\alpha=\infty$.

With regard to Mr. Gronner's introduction of the factor $\sqrt{1-v^2/c^2}$, this appears to be an *ad hoc* assumption without good authority. Admittedly, it is used freely in relativity analysis in many ways, but not to my knowledge as a factor to relate forces in different frames of reference as he has used it. It would appear that Mr. Gronner has introduced considerable novelty at this point. Moreover, if such a term is to be used at all to relate observed forces, I believe the manner of using it as proposed by Mr. Gronner will lead to results at variance with the law of conservation of momentum. This will be seen from the following analysis.

Conservation of momentum requires that what an observer notes as force must be proportional to the product of what he calls mass and the rate of change of velocity, regardless of the frame of reference. Since Mr. Gronner's problem is concerned only with components in the y direction, it is necessary to consider only the y components of force and acceleration. Taking the problem as he has set it up, and using his nomenclature, an observer in the stationary frame of reference will note that the force is

$$F_y=md^2y/dt^2 \tag{1}$$

while an observer in the moving frame of reference will note the force as

$$F_y^*=m^*d^2y^*/dt^{*2} \tag{2}$$

The problem is to find out whether Mr. Gronner's analysis yields this invariant relationship. That is, will an observer in the stationary system agree that equation 2 is correct for an observer in the moving system? Following Mr. Gronner

$\beta=\sqrt{1-v^2/c^2}$

$F_y=\beta F_y^*$

$y=y^*$ and therefore $d^2y=d^2y^*$

$t^*=(t-vx/c)/\beta$, and therefore $dt^{*2}=dt^2/\beta^2$ and $dt^2=\beta^2dt^{*2}$, a relationship which is mathematically sound and has the proper operational effect in the term d^2y/dt^2

It also commonly is accepted in relativity circles that

$m^*=m/\beta$, or $m=\beta m^*$

Substituting these values in equation 1 gives

$\beta F^*=\beta m^*d^2y^*/\beta^2dt^{*2}$

$$\text{or } F^*=m^*d^2y^*/\beta^2dt^{*2} \tag{3}$$

It is clear that equation 3 does not agree with equation 2, and the observer in one system will not agree with the observer in the other system. It would appear that if the factor $\sqrt{1-v^2/c^2}$ is to be used at all, Mr. Gronner's equation for the relationship between F_y and F_y^* should be $F_y = F_y^*/\beta$ instead of $F_y = \beta F^*$. But if this

should be the case, the term $(1-v^2/c^2)$ in his equation 16 would vanish, and the subsequent value of F_{2y} would be zero, as predicted in my article. It thus is seen that, if Mr. Gronner's result is accepted, one must discard the principle of the conservation of momentum, whereas if this principle is to be retained, one cannot accept Mr. Gronner's result.

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Sign of Reactive Power

To the Editor:

The article, "Sign of Reactive Power," which appeared in the November 1946 issue of *Electrical Engineering,* pages 512–16, is a most excellent presentation of the considerations which led a special subcommittee of the AIEE Standards committee to recommend that the present internationally recognized convention for the algebraic sign of reactive power should be reversed. The presentation is accurate and shows conclusively that, although either convention is possible mathematically, the present standard is the one which yields a vector power diagram which is similar to the admittance diagram and which is therefore the more useful for handling problems involving many individual loads connected in parallel. In spite of this result the subcommittee chose the opposite convention, because they believed that otherwise they logically would be compelled to abandon the even more useful concept that an "inductive load absorbs (positive) reactive power."

I believe, however, that their statement "no workable compromise which would allow the standard to remain unaltered and at the same time would meet the objections to it could be found" is unduly pessimistic. It seems to me that there is an alternative which will enable us to "eat our cake and have it, too," and which will cause the resulting minus sign to appear at a place where it will not do serious violence to recognized mathematical conventions.

The problem is to reconcile two desiderata, each of which is based on a long-recognized and significant fact, but between which there exists a certain implicit incompatibility. The first is

A. An overwhelming majority of the a-c circuits dealt with in modern electrical engineering are of the constant-voltage parallel-load type, and hence it is desirable to analyze them by vector diagrams for power which are similar to the vector diagrams of admittance and of current.

It was this consideration which led the AIEE committee on this subject in 1932 and the International Electrotechnical Commission in 1935 to select the present sign for reactive power. The second and equally significant fact is

B. An overwhelming majority of loads happen to be inductive, and it is therefore desirable that in metering and load dispatching the terminology should be consistent with the flow of a positive quantity analogous to energy from an overexcited generator to an induction motor.

This concept, that there is something which flows into an inductive load while reactive power is being maintained at its terminals, has become of great practical value. Var-hour meters have been invented to meter this imaginary quantity, and varmeters have been invented to indicate its rate of flow. Yet hitherto there has been no name for the quantity itself. This lack of adequate nomenclature has led to the unfortunate practice of speaking about "flow of reactive power," "absorption of reactive power," or for short "flow of reactive" and even "flow of *x*." Of course, strictly speaking reactive power does *not* flow but merely exists as a measure of the rate of flow of something else, just as active power does not flow but exists as a measure of the rate of flow of energy. Yet it is this concept, that something flows and that whatever does flow should be defined as positive, combined with the false notion that it is the reactive power which really flows, which has led to the recent request that the sign of reactive power be reversed.

I would like to propose as an alternative the following three-point program which seems to me to yield a more satisfactory solution than that offered by the subcommittee.

Point 1 is to coin a word to serve as a name for the quantity which is thought of as flowing into an inductive circuit and of which the var-hour is the unit. I propose the word, "quadergy" derived, of course, by combining syllables of "quadrature" and of "energy" (the term, "reactive energy" should not be used but should be reserved for the actual energy which alternately is stored and returned by an inductive circuit. This latter quantity is true energy, though it seldom is realized how small it is. For instance a circuit at which the reactive power is maintained at 1,000 kilovars stores and returns an amount of energy, the maximum value of which is only 0.0007 kilowatt-hours). The proposed definition is:

Quadergy 05.21.051

The quadergy which has been supplied by a source to a load during a time interval is the integral with respect to time of the reactive power at the points of entry of the source, taken over the time interval.

The value of quadergy is given in var-hours when the reactive power is in vars and the time is in hours.

Note: In describing the operation of an a-c system with reactive loads, it has been found convenient to assume the existence of this quantity, which is analogous to energy, which is inherently positive, and which is assumed to flow out of a properly excited source into an inductive load.

C. It follows from definition 05.21.050 that the quadergy supplied to an inductive load is positive because the reactive power at the points of entry of a source of quadergy is positive, while that at a sink of quadergy is negative.

Point 2 is to set up a standard convention for marking the scales and terminals of varmeters and related apparatus, so that there may be uniformity of practice among different power companies, and so that the instruments will indicate conveniently the flow of quadergy in the way most desired by load dispatchers. Presumably this will involve marking based on an "in and out" flow of quadergy rather than on "plus or minus" signs of reactive power or on "lag or lead" of current.

At its meeting in the spring of 1946 the AIEE committee on instruments and measurements authorized the formation of a subcommittee for this purpose.

Point 3 is to leave the present definition of the sign of reactive power unchanged.

It will be seen that Point 2 satisfies the desideratum *B*, and Point 3 satisfies the desideratum *A*, and that Point 1 merely gives recognition to and makes more convenient the use of a concept which long experience has shown to be of great value in the power industry.

With this procedure the incompatibility will show up in a difference in the practice of connecting varmeters and wattmeters. This difference can be expressed in either of two ways.

1. If we still retain the arbitrary convention that an upscale deflection of an instrument is *always* to correspond to a positive value of the quantity (active power or reactive power) indicated by the instrument, then it appears that at a given set of points of entry, which divide a complete circuit into two portions, the wattmeter indicates the active power of one portion while the varmeter indicates the reactive power of the other portion. The magnitude of either power is necessarily the same for both portions of the circuit.

2. The alternative and simpler way of describing this situation is to say that we scrap the arbitrary convention which connects the direction of an instrument deflection and the sign of the indicated quantity

Viewed on either alternative this situation is closely parallel to, and no worse than, that of a center-zero ammeter in a storage battery circuit. Here custom requires that an upscale deflection correspond to a condition of charge. If the ammeter is in the positive lead of the battery, this upscale deflection does correspond to a positive current, but, if the ammeter happens to be connected in the negative lead, either (1) the ammeter must be considered as measuring the current in the charging circuit instead of the current in the battery or else (2) the upscale deflection on charge must be taken to correspond to a negative current, that is, a flow of electric charge out of the load.

If the actual connection of wattmeters and varmeters on the switchboard were made directly on the basis of fundamental definitions, this situation might be objectionable. In practice, however, the elec-

trician wires the instrument in accordance with the manufacturer's blueprint. The blueprint is based on current practice for wattmeters and on the standard as set up in Point 2 for varmeters, and I believe no difficulties will be created. In fact, at present the generator wattmeters are connected to read upscale for energy flow *out* of the generator, while motor wattmeters are connected to read upscale for energy flow *into* the motor. I propose that we similarly connect generator varmeters to read upscale when quadergy is flowing *out* of the generator and connect motor varmeters so as to read upscale when quadergy is flowing *into* the motor. This ignores the fact that at the generator the active power is negative while the reactive power is positive, because I believe that this inconsistency is much less disturbing than an abandonment of desideratum *A*.

Quadergy is of course a highly artificial and intangible concept, yet it would seem that definite progress should result if engineers should give to this "airy nothing, a local habitation and a name."

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NEW BOOKS....

"Heating Ventilating Air Conditioning Guide 1946." This 24th edition of the guide contains an enlarged technical data section and a new grouping of subjects "to provide a more logical progression from fundamental principles to specific application." A new chapter on "Fluid Flow" has been added, and important additions and revisions have been made in the chapters: Thermodynamics, Air Contaminants, Instruments and Measurements, Heat Transmission Coefficients of Building Materials, Physiological Principles, Air Conditioning in the Prevention and Treatment of Disease, Cooling Load, Heating Boilers and Furnaces, Gravity Warm Air Systems, Mechanical Warm Air Systems, Steam Heating Systems and Piping, Pipe Insulation, Dehumidification by Sorbent Materials, Refrigeration, Air Distribution, Air Duct Design, and Owning and Operating Costs. The chapter on terminology has been revised to obtain better conformity with the terms of basic science rather than the less accurate forms often adopted by specialists. Besides the 51-chapter technical data section, a manufacturers' catalogue data section, and the roll of membership of the American Society of Heating and Ventilating Engineers are included in the volume. Complete indexes to the technical and catalogue sections are provided. The technical data section is divided into eight sections: princspals, human reaction to atmospheric environment, heating and cooling loads, combustion and consumption of fuel, heating systems and equipment, air conditioning, special applications, installation and testing codes. The American Society for Heating and Ventilating Engineers, New York 10, N. Y., 1946, $6^1/_4$ by $9^1/_4$ inches, 1,279 pages plus the roll of membership, cloth, $6.

The following new books are among those recently received at the Engineering Societies Library. Unless otherwise specified, books listed have been presented by the publishers. The Institute assumes no responsibility for statements made in the following summaries, information for which is taken from the prefaces of the books in question.

ASTM STANDARDS ON RUBBER PRODUCTS. Prepared by ASTM Committee D-11 on Rubber Products; Methods of Testing, Specifications, March, 1946. American Society for Testing Materials, Philadelphia 2, Pa. 540 pages, illustrated, 9 by 6 inches, **paper, $3.25 (to ASTM members, $2.20).** More than 75 specifications and tests covering natural and synthetic rubbers are collected in this publication. The largest section covers general methods of analysis, identification, and testing. The following sections deal with electric tests; insulated wire and cable; hose, belting, gloves, tape, and so forth, nonrigid plastics; and tests and specifications for miscellaneous products not previously covered. There is a section on nomenclature and definitions, and a number of proposed specifications are included to solicit comment.

CIRCUIT ANALYSIS. By C. E. Skroder and M. S. Helm. Prentice-Hall, Inc., New York, N. Y., 1946. 288 pages, illustrated, $9^1/_4$ by 6 inches, cloth, $5.35. The object of this book is to provide in a single volume a wide selection of laboratory circuit experiments, including a very comprehensive coverage of the related circuit theory and laws, the theory and limitations of the necessary instruments, and the methods of making measurements. The experiments are presented as problems to induce the student to adopt an analytical approach in determining the experimental data needed, the circuits to be used, the measurements to be made, and the interpretation of the observed and calculated data for a rational solution.

CURRENTS IN AERIALS AND HIGH-FREQUENCY NETWORKS. By F. B. Pidduck. Clarendon Press, Oxford, England; Oxford University Press, New York, N. Y., 1946. 97 pages, illustrated, 9 by $5^1/_2$ inches, cloth, $2.50. This book is an account of previously unpublished investigations on currents in aerials, based on a little-known paper of Pocklington. Fundamental formulas are deduced and extended to soldered networks. The theory of transmission lines is established as that of a system of two aerials, the untuned Y feeder is considered in detail, and a theory of aerails parallel to the earth is worked out. The necessary tables for calculations are included.

INDUCTANCE CALCULATIONS. By F. W. Grover. D. Van Nostrand Company, New York, N. Y., 1946. 286 pages, illustrated, $9^1/_4$ by 6 inches, cloth, $5.75. This volume has been prepared with the idea of providing for each special type of inductor a single simple formula that will involve only the parameters that naturally enter together with numerical factors obtainable from tables. General principles and working methods are given in the introductory chapters. The separate types of circuits are covered in the succeeding chapters, including the necessary tables with the last three chapters devoted to auxiliary table of functions and special formulas.

PLASTICS FOR ELECTRICAL AND RADIO ENGINEERS. By W. J. Tucker and R. S. Roberts. Technical Press, Ltd., Gloucester Road, Kingston Hill, Surrey, England, 1946. 148 pages, illustrated, 9 by $5^1/_4$ inches, cloth, 12*s*. This handbook provides the electronic engineer with essential data relating to the application of plastics in the electrical and radio industry. Molding and manufacturing procedure and the advantages and limitations of plastic materials are discussed in a full and practical manner. Insulation and testing problems are dealt with in detail, and tables giving physical properties are presented in a manner useful to the electronic industry.

CALCULUS. By F. H. Miller. Second edition. John Wiley and Sons, New York, N. Y.; Chapman and Hall, London, England, 1946. 416 pages, illustrated, $8^1/_2$ by $5^1/_2$ inches, cloth, $3.50. This book is designed to give the student a comprehension of the basic concepts and methods of calculus, presenting the subject both as an important branch of mathematics and as a tool for practical use. Additions to the new edition include a discussion of graphical differentiation, a summary of the processes of integration, a discussion of approximate integration, and numberous formulas and theorems. Answers are given to the odd-numbered exercises only, the balance of the answers being available spearately to teachers.

PIEZOELECTRICITY. By W. G. Cady. McGraw-Hill Book Company, New York, N. Y., 1946. 806 pages, illustrated, 9 by $5^3/_4$ inches, cloth, $9. The entire field of piezoelectricity is covered in this comprehensive treatise, including related areas of elasticity, dielectrics, optics, and magnetism. Crystallography and the general properties of crystals lead up to the discussion of quartz and Rochelle salt with their special applications. A unified account is given of experimental results, with many formulas, numerical data, and an extensive bibliography.

LES TENSEURS EN MÉCANIQUE ET EN ELASTICITÉ. By L. Brillouin. Dover Publications, New York, N. Y., 1946. 364 pages, illustrated, $9^1/_2$ by 6 inches, cloth, $3.75. This book presents an exposition of tensor analysis and its applications in theoretical physics. The range of the book is indicated by the chapter headings: vector geometry; pseudo-tensors; the principal differential operators; parallelism, covariant, and contravariant derivative; Riemann space; use of Riemannian geometries in analytical mechanics; transition of wave mechanics; elasticity and waves in elastic solids; theory of solids and quantum mechanics. There is an index.

LES RADIATIONS. By C. Fabry. Librairie Armand Colin, Paris, France, 1946. 220 pages, illustrated, $6^1/_4$ by $4^1/_2$ inches, paper, 60 francs. The fundamental properties of radiations, their sources, and effects are discussed in this small volume. The author presents a critical study of recent developments in producing, analyzing, and measuring thermal and optical radiations, including fluorescence and phosphorescence. The final chapter describes biologic and therapeutic applications and uses in analytic chemistry and photochemistry.

LABOR PROBLEMS. By W. V. Owen. Ronald Press Company, New York, N. Y., 1946. 570 pages, illustrated, $9^1/_4$ by 6 inches, cloth, $4.50. Intended to present a survey of the principles and problems of labor economics from a broad viewpoint, this book covers the subject in four main sections: the economic characteristics of labor and the nature of the labor market; the labor market in operation, including supply and demand, wages, hours, working conditions, and the mobility of labor; the control of the labor market exerted by the trade unions, the employer, and the state; and security in the labor market as affected by old age, unemployment, and time lost through strikes and disagreements.

PSYCHOLOGY IN INDUSTRY. By N. R. F. Maier. Houghton Mifflin Company, Boston, Mass., New York, N. Y., and Chicago, Ill., 1946. 463 pages, illustrated, $8^1/_4$ by $5^1/_4$ inches, cloth, $3. The author discusses in a nontechnical but systematic manner the various aspects of the industrial situation to which the principles of psychology are applicable. Beginning with the fundamental problem of why people act as they do, he covers morale and motivations, the allowance for variations in individuals, the use of psychological tests and the analysis of skills, fatigue conditions and the prevention of accidents, and psychological factors in labor turnover. The book is designed for a wide audience from the student to top management.

RADAR. By J. F. Rider and G. C. B. Rowe. John F. Rider, Publisher, Inc., New York 16, N. Y., 1946. 72 pages, illustrated, 11 by $8^1/_2$ inches, paper, $1. The underlying principles of radar are explained in simple language with a description of the basic radar set. The use of radar during the war by land, sea, and air forces is described, including countermeasures. A brief survey of the future of radar is added in conclusion.

STEAM POWER PLANT AUXILIARIES AND ACCESSORIES. By T. Croft, editor, revised by D. J. Duffin, 2nd edition. McGraw-Hill Book Company, Inc., New York, N. Y., and London, England, 1946. 583 pages, illustrated, $8^1/_4$ by $5^1/_2$ inches, cloth, $5. A practical manual for the operating engineer, this book has been revised to conform with the changes that have taken place in the 24 years since it was published originally. Topics on which considerable new material has been added include reciprocating and centrifugal pumps, methods of boiler feeding, feed-water heaters, economizers and air preheaters, condensers, steam piping of power plants, and steam traps. As before, there are review questions and problems, and a technical data section has been added.

ELECTRON MICROSCOPE. By E. F. Burton and W. H. Kohl. Second edition. Reinhold Publishing Corporation, New York, N. Y., 1946. 325 pages, illustrated, $9^1/_4$ by 6 inches, cloth, $4. The opening chapters cover the principles of the optical microscope, including a discussion of wave motion as a means of propagation of energy. The author proceeds to a description of electromagnetics and the electron, and then presents the physical principles upon which the operation of the electron microscope is based. Both the electrostatic and magnetic types are covered. Important practical applications of the electron microscope are considered in the last chapter. The book is well illustrated with sketches and microphotographs, and there is a 20-page bibliography.

GRUNDZÜGE DER THEORETISCHEN LOGIK. By D. Hilbert and W. Ackermann. Second enlarged edition. Dover Publications, New York, N. Y., 1946. 133 pages, tables, $8^3/_4$ by $5^1/_2$ inches, cloth, $2.50. An introduction to the subject of symbolic logic, this book demonstrates the application of strict formulas and methods for their manipulation to the field of philosophical logic. Four chapters deal respectively with: the propositional calculus; the calculus of classes; the simple function calculus, that is, the calculus of classes of individuals and of multiple relations between individuals; and the logical paradoxes and the use of a theory of type to avoid paradox.

MACRAE'S BLUE BOOK. Fifty-third annual edition. MacRae's Blue Book Company, 18 East Huron Street, Chicago, Ill., 1946. 3,736 pages, illustrated, 11 by 8 inches, cloth, $15. This annual directory gives a complete listing of manufacturers in the United States, classified according to products. The first 400 pages of the volume contain an alphabetical list of manufacturers, producers and wholesalers with capital ratings, location of branches, and, in some cases, addresses of local distributors. The third and last section of the volume is an alphabetical list of trade names.

ELECTRIC DISCHARGE LAMPS. By H. Cotton. Chapman and Hall, London, England, 1946. 435 pages, illustrated, $8^3/_4$ by $5^1/_2$ inches, cloth, 36*s*. The structure of atoms and molecules is explained in sufficient detail for an understanding of the fundamental principles involved in light production in discharge lamps. Descriptions are given of the various kinds of lamps in present use, with their construction and operating characteristics. The control of electric discharge lamps is discussed, and a chapter is devoted to the nature of fluorescence and the applications of fluorescent materials to these lamps.

ENGLISH-FRENCH AND FRENCH-ENGLISH TECHNICAL DICTIONARY. By F. Cusset. Chemical Publishing Company, Brooklyn, N. Y., 1946. 590 pages, $6^3/_4$ by 5 inches, cloth, $5. Giving both English to French and French to English translations, this dictionary, as shown on the title page, covers metallurgy, mining, electricity, chemistry mechanics, and science. Phrases as well as words are given and in many cases appear directly under each of the important words. A few basic conversion tables appear at the end of the book.

LABOR-MANAGEMENT ECONOMICS. By W. V. Owen in collaboration with Stevenson, Jordan and Harrison, Inc., Management Engineers. Ronald Press Company, New York, N. Y., 1946. 121 pages, diagrams, $8^1/_4$ by $5^1/_4$ inches, cloth, $2. Part I, on employer-employee economics, deals with management, production, risks, and costs on the employer's side of the question and with wages, labor economics, and unionism, as more directly related to the employees' interests. It ends with a chapter on industrial relations in general. Part II, the economic framework, is concerned with some of the economic forces, concepts, and relations that provide the economic mechanics for producing, marketing, and consuming goods.

ELECTRONICS IN INDUSTRY. By G. M. Chute. McGraw-Hill Book Company, New York, N. Y., and London, England, 1946. 461 pages, illustrated, $8^1/_2$ by $5^1/_4$ inches, cloth, $5. To give a broad introduction to the use of electronic circuits and equipment is the purpose of this book. It outlines the industrial uses of tube circuits and gives detailed explanation of a large number of electronic equipments now serving in industrial plants. No previous knowledge of tubes is assumed, the early chapters being devoted to the necessary fundamentals, and, since the book is intended for users of equipment already available, no design information is presented.

ATOMIC SPECTRA. By R. C. Johnson. Methuen and Company, Ltd., London, England. Distributed by Chemical Publishing Company, Brooklyn, N. Y., 1946. 120 pages, illustrated, $6^3/_4$ by 4 inches, cloth, $2. Following a brief introduction on the production of spectra, the book discusses spectra of the hydrogen type, one electron revolving around a positive nucleus, and spectra of atoms with two or more electrons. The structure of atoms is considered, and two chapters deal with the effect of special conditions and applied fields on spectral lines. The organization of the periodic table on the basis of the electronic structure of the elements is demonstrated, and spectroscopic instruments and procedures are described.

PRINCIPLES CF PHYSICS. Volume 2, Electricity and Magnetism. By F. W. Sears. Addison-Wesley Press, Cambridge 42, Mass., 1946. 434 pages, illustrated, 9 by 6 inches, cloth, $5. A college textbook, this volume deals with the elements of electricity and magnetism on the basis of a thorough preliminary background of mathematics and general physics. Separate chapters are devoted to ferromagnetism, electromotive force, and the basic principles of electronics.

FUNDAMENTALS OF ALTERNATING-CURRENT MACHINES. By A. Pen-Tung Sah. McGraw-Hill Book Company, New York, N. Y., and London, England, 1946. 466 pages, illustrated, 9 by $5^3/_4$ inches, cloth, $5. Written from the viewpoint of the operating man rather than that of the design engineer, this book lays particular stress on the operating problems of electric machines. The student is shown not only how to test his equipment, but also what tests to make in order to obtain quickly and accurately the requisite parameters from which he may compute the performance. Thus design formulas are avoided, while the experimental determination of the machine parameters is given full prominence after the theory is explained. Only those theories and experimental methods capable of being generalized are treated, and for directness the theory is developed from the standpoint of circuits.

WAVE PROPAGATION IN PERIODIC STRUCTURES. By L. Brillouin. McGraw-Hill Book Company, Inc., New York, N. Y., and London, England, 1946. 247 pages, illustrated, $8^1/_2$ by $5^1/_4$ inches, cloth, $4. Based on lectures given at the University of Wisconsin, this volume includes a variety of problems having a common mathematical background. They extend from electrical engineering to electromagnetism and wave mechanics of the spinning electron. The book includes explanations of electric filters, rest rays, anomalous optical reflection, selective reflection of X rays or electrons from a crystal, and omission of energy dissipation.

UNDERSTANDING MICROWAVES. By V. J. Young. John F. Rider Publisher, New York 16, N. Y., 1946. 385 pages, illustrated, $8^3/_4$ by $5^1/_2$ inches, cloth, $6. This book explains the fundamental problems encountered in the field of ultrahigh frequency research and production and how they are surmounted—problems met in the design and operation of wave guides and coaxial lines; resonant cavities as they function in the magnetron, the dynatron, and the klystron; the theory and design of antennas used in conjunction with the transmission and reception of microwaves. The underlying electromagnetic and electrostatic theories are interpreted in the early chapters. Section II contains a comprehensive collection of terms, ideas, and theorems with thorough definitions and explanations.

TRIGONOMETRY. By H. K. Hughes and G. T. Miller. Second edition. John Wiley and Sons, New York, N. Y.; Chapman and Hall, London, England, 1946. 175 pages, illustrated, $8^3/_4$ by $5^1/_2$ inches, cloth, $2.50. Intended for use in a first course in trigonometry for college freshmen, this book aims particularly at clearness and ease of understanding. To this end, certain changes in the order and amount of material have been made in the new edition. More than customary attention is given to the law of cosines and to the use of logarithms with this law. No tables are included in the volume.

PHYSICAL AND CHEMICAL EXAMINATION OF PAINTS, VARNISHES, LACQUERS, AND COLORS. By H. A. Gardner and G. G. Sward, Tenth edition. Henry A. Gardner Laboratory, 4723 Elm Street, Bethesda, Md., 1946. 652 pages, illustrated, 12 by $10^1/_4$ inches, fabrikoid, $18.50. The most comprehensive work in its field, this revised edition presents current standard test methods and experimental methods, as well as some that may temporarily be obsolete. It includes charts of color families and also charts of colors used by the Army and the Navy. The book covers all types of examination used in its field. It is well illustrated and has many charts and diagrams.

ORGANIZED LABOR AND PRODUCTION. By M. L. Cooke and P. Murray. Revised edition. Harper and Brothers, New York, N. Y., and London, England, 1946. 277 pages, illustrated, $8^3/_4$ by 6 inches, cloth, $2.50. A prominent expert in scientific management and the head of a leading union join in setting forth what they both believe is a basis for greater collaboration among employers and organized workers to assure greater production under fair terms of employment. The book covers the responsibilities and activities of both labor and management and discusses some of the questionable practices of both.

METALLIC CORROSION, PASSIVITY AND PROTECTION. By U. R. Evans. Appendix by A. B. Winterbottom. Longmans, Green and Company, New York, N. Y., 1946. Revised edition, 863 pages, illustrated, $9^1/_2$ by $5^3/_4$ inches, cloth, $14. The material in the new edition has been rearranged with considerable addition of recent developments. The first two chapters present simple examples of corrosion and passivity and a study of thin films. Corrosion under varying conditions, such as with and without oxygen, is described in succeeding chapters, followed by a discussion of the influence of stress, structure, crevices, and the like. Four chapters are devoted to protection by various methods, and the book ends with a section on testing. Each chapter is divided into three parts: the scientific basis; the technical aspects including practical problems; and the mathematical, or quantitative, treatment.

PAMPHLETS . . .

The following recently issued pamphlets may be of interest to readers of "Electrical Engineering." All inquiries should be addressed to the issuers.

How to Use Carbon Tetrachloride Safely: Tips to the Foreman. Safety Research Institute, Inc., 420 Lexington Avenue, New York 17, N. Y., single copies, five cents. Quantity prices will be quoted.

The Action of Embeco in Concrete and Mortars (second edition). The Master Builders Company, 7016 Euclid Avenue, Cleveland 3, Ohio, 34 pages, no charge.

Celanese Synthetics for the Electrical Industry, Celanese Plastics Corporation, 180 Madison Avenue, New York 16, N. Y., 20 pages, no charge.

The Influence of the Concentration and Mobility of Ions on Dielectric Loss of Insulating Oils

BUN PO KANG
ASSOCIATE AIEE

Synopsis: The causes of dielectric loss in insulating oils and the means to reduce it to a minimum have always been important in high voltage engineering. In this study the relationships between the amount of impurities, which furnish a source of free ions in an oil, and the viscosity, which influences the mobility of the ions, on the one hand and the dielectric loss on the other are investigated. The results present some definite relationships between them and point out that in two oils of the same "electrical purity," the one with higher viscosity will have lower dielectric loss.

SINCE DIELECTRIC LOSS is a matter of primary importance in high voltage insulation, many phases of the phenomenon have been a subject of frequent discussion among high voltage engineers. In liquid dielectrics, such as mineral oils which are used extensively for transformer and cable insulations and for many other purposes, the loss has been found to be caused largely by ionic conduction.[1,2] In a few cases, dielectric absorption caused by molecular polarization has been found to be present also.[1] However, polarization in oils resulting in an absorptive component of dielectric loss seldom appears significant.

In the study of the nature of and the mechanism that causes dielectric loss in oil, where the loss is due to ionic conduction alone, there are two important factors which merit investigation. In order to have ionic conduction, the oil must possess ions, and the ions must be free to move. If the ion content of an oil is constant, the dielectric loss should vary inversely as the viscosity. It has been found that within a limited rage of temperature, this relation holds true, and the product of the conductivity and the viscosity of the oil is approximately a constant. This product has been suggested as a measure of the free ion content, which has also been termed the "electrical purity" of the liquid.[3]

Under constant electric stress, the dielectric loss for liquids without significant dielectric absorption is proportional to the conductivity of the liquid.

$$W=\frac{\lambda A E^2}{L} \tag{1}$$

where

A is the area of electrode in square centimeters
L is the electrode separation in centimeters
E is the applied potential in volts
λ is the conductivity in mhos per centimeter
W is the dielectric loss in watts

Since in the cases of ionic conduction the conductivity should be directly proportional to the ion content, it follows that the ion content also should be proportional to the product of the viscosity and the dielectric loss. However, this is not always borne out by experiment except within a very small range of temperature. If the range of temperature be extended, the ion content is found to increase with temperature. Apparently this is caused by thermo-agitation, which in turn causes further ionic dissociation and the generation of new ions within the oil.[4]

For the same degree of purification, the dielectric loss is invariably lower in an oil of higher viscosity. Even though the amount of impurity and the number of ions present in two different oils may be identical, the mobility of the ions is much greater in the oil of lower viscosity and consequently the conduction current is higher.

Since it is not only interesting but useful to know the relationships between the concentration and mobility of the ions of an insulating oil on the one hand and dielectric loss on the other, it is the purpose of this work to establish a correlation between them.

The Experiment

Three original oils, designated as oil sample 1, oil sample 6, and oil sample 7, were used. Samples 1 and 7 were highly refined oils furnished by well-known refineries, sample 1 being of low viscosity and sample 7 being of high viscosity. Sample 6 was an oil known to be highly susceptible to oxidation and was subjected to severe oxidation at 150 degrees centigrade under atmospheric condition for seven days. Oil samples 2 to 5 were mixtures of samples 1 and 6 (by weight) at 10, 25, 50, and 75 per cent, respectively, of oil 6. Oil samples 8 to 11 were mixtures of oils 7 and 6 in similar percentages.

For the measurements of viscosity and dielectric loss, these samples were classified into two groups. Group A, oil samples 1 to 6, consisted of the low viscosity samples, and group B, oil samples 7 to 12, consisted of the high viscosity samples. Samples 6 and 12 were the same oil.

Viscosity measurements were made with a standard Saybolt viscosimeter. The 60-cycle power factor measurements were made with an Atkinson bridge.[5] The samples were measured in a nickel cell with an electrode area of 40.742 square centimeters and a distance of separation of 1.27 millimeters at a voltage stress of 197 volts per millimeter. The values of dielectric loss given on the following figures are the total loss of each specimen under test. If the loss per centimeter cube is desired, these values

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Table I. Viscosities of the Oil Samples at Different Temperatures

Sample	40 C	60 C	80 C	100 C	120 C	150 C
1	—	0.086	0.05	0.031	0.02	—
2	0.21	0.10	0.055	0.035	0.021	—
3	0.305	0.13	0.073	0.046	0.03	—
4	0.65	0.23	0.115	0.065	0.04	—
5	1.60	0.53	0.205	0.10	0.06	—
6	4.90	1.10	0.40	0.17	0.094	0.048
7	5.00	1.40	0.54	0.24	0.14	0.074
8	—	1.40	0.54	0.24	0.135	0.069
9	—	1.35	0.505	0.23	0.13	0.066
10	—	1.35	0.455	0.201	0.105	0.06
11	—	1.20	0.43	0.19	0.10	0.059
12	4.90	1.10	0.40	0.17	0.094	0.048

The values of viscosity given are in poises.

Table II. Power Factor and Dielectric Loss of Oil Samples in Group *A*

Sample		40 C	60 C	80 C	100 C	120 C	150 C
1	Capacitance	—	66.7	66.1	65.7	65.0	—
1	Power factor	—	0.18	0.41	0.61	1.56	—
1	Dielectric loss	—	2.9	6.3	12.6	23.9	—
2	Capacitance	67.7	67.2	66.7	66.0	65.5	—
2	Power factor	0.55	1.59	4.39	10.20	18.25	—
2	Dielectric loss	8.82	25.2	69.3	167.5	283.0	—
3	Capacitance	68.6	67.9	67.5	66.7	66.0	—
3	Power factor	0.82	2.58	7.10	17.2	33.8	—
3	Dielectric loss	13.2	41.5	113.0	272.0	530.0	—
4	Capacitance	69.4	68.8	68.5	67.6	67.2	—
4	Power factor	0.75	2.72	7.85	20.1	36.3	—
4	Dielectric loss	11.95	44.1	127.5	322.0	575.0	—
5	Capacitance	70.7	70.0	69.4	69.0	68.5	—
5	Power factor	0.36	1.85	6.52	19.7	39.9	—
5	Dielectric loss	5.97	30.2	107.0	322.0	642.0	—
6	Capacitance	71.8	71.0	70.5	69.5	68.8	68.1
6	Power factor	0.19	0.98	4.55	15.7	42.5	89.8
6	Dielectric loss	3.75	16.4	75.6	258.0	693.0	1450.0

Capacitance is given in micromicrofarads, power factor in per cents, and dielectric loss in microwatts.

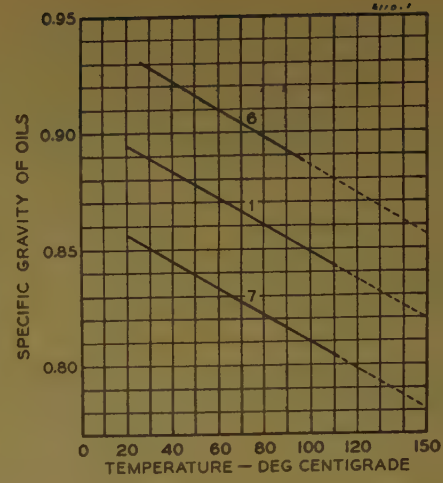

Figure 1. The variation of specific gravity with temperature of the original oil samples

must be multiplied by a factor of 0.00312, which is the ratio of L/A, where L is the distance of separation and A is the effective area. The voltage stress was kept low to utilize the linear portion of the well-known current-voltage relation curve. For group A, the low viscosity oils, readings were taken at 40, 60, 80, 100, and 120 degrees centigrade for each sample, and for group B, the high viscosity group, readings were taken at 60, 80, 100, 120, and 150 degrees centigrade.

Experimental Results

The specific gravity temperature relations of the three original oil samples are given in Figure 1. The specific gravity of each of the mixtures at various temperatures can be computed easily from these curves by the use of the following relations:

$$\rho_x = \frac{\rho_1 \rho_2 (W_1 + W_2)}{\rho_1 W_2 + \rho_2 W_1} \quad (2)$$

where

ρ_1, ρ_2 represent the specific gravities of the components

ρ_x represents the specific gravities of their mixture

W_1, W_2 represent the weights of the components

After converting the viscosity readings at different temperatures for each sample from Saybolt seconds into poises, the results were given in Table I. To follow the more customary way of representing relationships of this kind, the logarithm of viscosity was plotted against the reciprocal of the absolute temperature in Figures 2 and 3. It is quite obvious that the relationship is essentially logarithmic. However, none of the curves is exactly a straight line but only very near to one.

The values of power factor and dielectric loss for each mixture at different temperatures are given in Table II for group A, and Table III for group B. Figure 4 gives the relations of the logarithm of dielectric loss as a function of the reciprocal of absolute temperature. Several of these curves are essentially straight lines and others are quite near to straight lines within limited range of temperatures. These relations may be expressed as

$$W = A - \frac{B}{T} \quad (3)$$

where

W is dielectric loss in microwatts
T is temperature in degrees Kelvin
A and B are constants

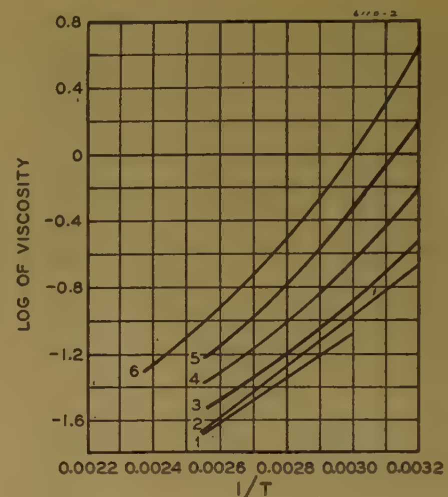

Figure 2. The viscosity–temperature characteristics of the oil samples in group A, η being viscosities in poises, T being temperatures in degrees Kelvin, and numbers designating the oil samples

which are similar to the expression used by Fuoss for the relation between conductivity and temperature.[6]

Dielectric Loss and Viscosity Relation

Based on the curves of Figures 2, 3, and 4, a relationship between viscosity and dielectric loss of each sample can be established. Figure 5 presents clear evidence that the relation is logarithmic,

$$\log W = A - B \log \eta \quad (4)$$

The curves are approximately straight lines within reasonable allowance for experimental error, and each group of samples appears to have a constant slope of its own. These results confirm the prediction of other workers in this field.[4]

Discussion

Since the dielectric loss of sample 6 is much higher than that of samples 1 and 7 (from 6 to 30 times under similar conditions), the ions present in the various mixed samples may be assumed as coming from sample 6. If there were no other causes for further dissociation in the oil, the dielectric loss of each sample should be a linear function of the percentage of deteriorated oil added. Figure 6 shows dielectric loss as a function of percentage of deteriorated oil added at different constant temperatures. The curves of group A for temperatures below 100 degrees centigrade show an increase of dielectric loss with the percentage of deteriorated oil added, but only up to a certain point, and beyond this point the losses decrease again. The ascending portion of each curve is caused by the increase of ion content of the specimen by adding a larger quantity of deteriorated oil. Since the deteriorated oil, sample 6, has much higher viscosity than sample 1, upon addition of sample 6, the viscosity of the mixture increases rapidly. Even though the ion content continues to increase, the mobility of these ions is greatly reduced by the increase in viscosity. After reaching a certain point, the viscosity becomes so high that even though the ion content continues to increase, the mobility of the ions diminishes rapidly and the dielectric loss decreases again.

Based on these curves, if one assumes that the ion content N is equal to the product of viscosity η and dielectric loss W, mathematical analysis shows that in Figure 6 all curves in group A should pass through a maximum, but not the curves in group B. This is found to be true. However, because of the fact that sludges and other solid impurities might have been formed in oil 6 because of oxidation, the viscosities of the mixtures did not follow exactly the relation usually used for computing the viscosity of mixtures of nonpolar liquids.

$$\frac{1}{\eta_x}=\frac{X}{\eta_1}+\frac{1-X}{\eta_2} \tag{5}$$

where

η_1, η_2, η_x are viscosities of the components and the mixture

X is the percentage of the liquid with viscosity η_1

The theoretical values are lower in comparison with experimental data. Based on theoretical considerations, one can show that the maximums of dielectric loss occur when

$$X=\frac{\eta_1}{2(\eta_1-\eta_2)} \tag{6}$$

provided η_1 is not equal to η_2. The experimental values deviated from this prediction somewhat because of the high viscosities obtained. However, the general shapes of the curves support very well the assumption that

$$W=\frac{N}{\eta} \tag{7}$$

In group B, both samples 6 and 7 have approximately the same viscosity; in fact, the viscosity of sample 6 is a little lower than the viscosity of sample 7, and consequently the increase of the percentage of deteriorated oil increases both the concentration and mobility of the ions, and the conduction current and dielectric loss also are increased.

In a limited range of temperature within which new ions are not generated by thermo-agitation in the oil, the dielectric loss should be proportional to the percentage of deteriorated oil added if the viscosity of the mixture is kept constant. This is found to be true for samples of higher viscosities according to Figure 7, curves B and C. In this group the viscosities of the two components were only slightly different, and the temperature differences were small when the samples were brought to the same viscosity. Consequently, thermal effects may be considered negligible.

If the viscosities of the two components

Table III. Power Factor and Dielectric Loss of Oil Samples in Group B

Sample		60 C	80 C	100 C	120 C	150 C
7	Capacitance	64.0	63.2	62.8	62.4	61.1
7	Power factor	0.062	0.25	0.59	1.28	3.96
7	Dielectric loss	0.945	3.78	8.82	18.9	57.3
8	Capacitance	64.5	64.0	63.2	62.8	61.8
8	Power factor	0.25	0.58	1.51	3.05	9.0
8	Dielectric loss	3.78	8.82	22.7	45.3	132.48
9	Capacitance	65.5	65.2	64.5	63.8	63.0
9	Power factor	0.31	1.11	3.05	6.7	14.8
9	Dielectric loss	4.73	17.0	46.7	101.0	220.0
10	Capacitance	67.5	66.8	66.3	66.0	64.5
10	Power factor	0.56	1.99	6.01	13.3	25.5
10	Dielectric loss	8.82	31.5	94.5	208.0	392.0
11	Capacitance	69.3	68.6	67.9	67.2	67.1
11	Power factor	0.81	3.42	11.0	24.2	47.6
11	Dielectric loss	13.25	55.5	177.0	385.0	756.0
12	Capacitance	71.0	70.5	69.5	68.8	68.1
12	Power factor	0.98	4.55	15.7	42.5	89.8
12	Dielectric loss	16.4	75.6	258.0	693.0	1450.0

Capacitance is given in micromicrofarads, power factor in per cent, and dielectric loss in microwatts.

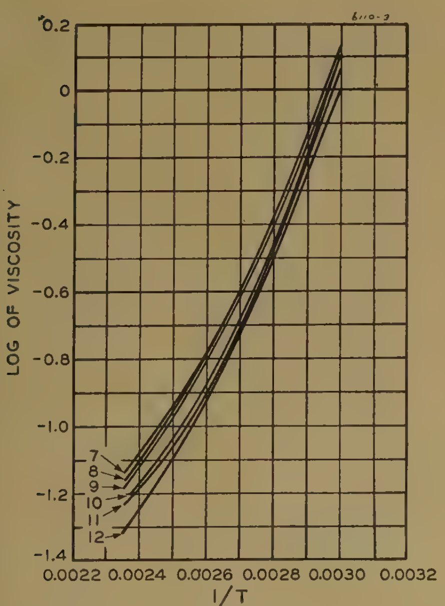

Figure 3. The viscosity–temperature characteristics of the oil samples in group B, η being viscosities in poises, T being temperatures in degrees Kelvin, and numbers designating the oil samples

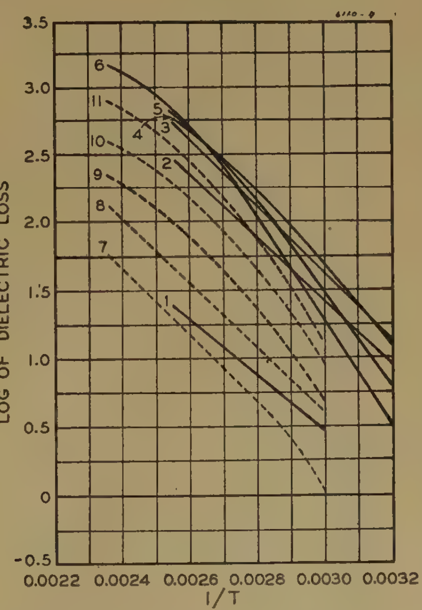

Figure 4. The variation of dielectric loss with temperature, loss being given in microwatts, temperature in degrees Kelvin, solid lines for oils in group A, and dotted lines for oils in group B

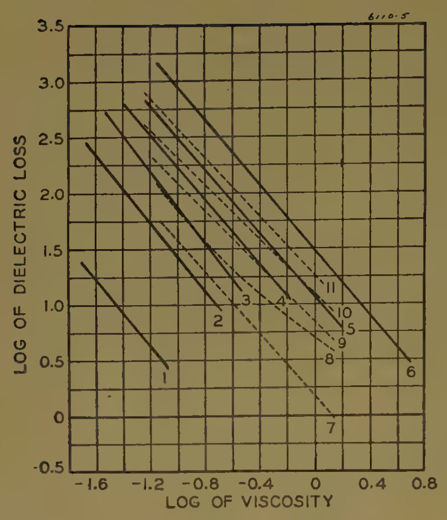

Figure 5. The variation of dielectric loss with viscosity, loss being given in microwatts and viscosity in poises

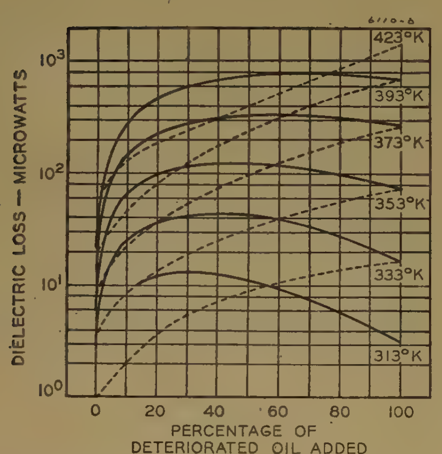

Figure 6. The variation of dielectric loss with percentage of deteriorated oil added at various temperatures, solid lines representing oils in group *A* and dotted lines, oils in group *B*

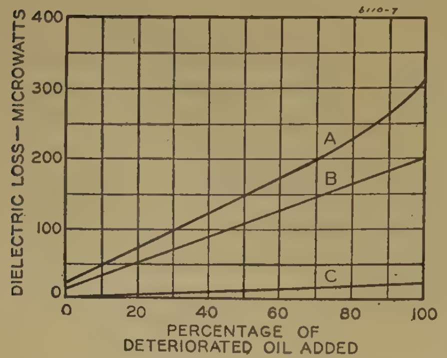

Figure 7. The influence of deteriorated oil on dielectric loss at constant viscosity for high viscosity mixtures

Curve *A* at 0.15 poise, curve *B* at 0.2 poise, and curve *C* at 1.0 poise

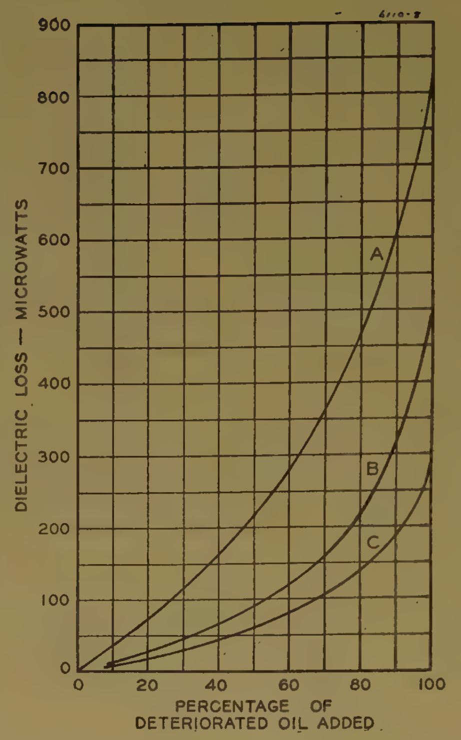

Figure 8. The influence of deteriorated oil on dielectric loss at constant viscosity for low viscosity mixtures

Curve *A* at 0.08 poise, curve *B* at 0.15 poise, and curve *C* at 0.2 poise

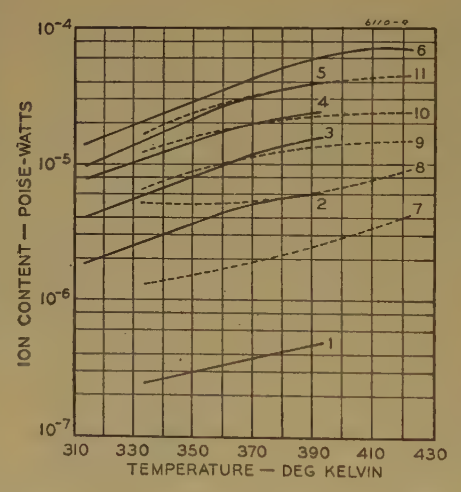

Figure 9. The variation of ion content with temperature, solid lines representing oils in group *A* and dotted lines, oils in group *B*

of the mixture are greatly different, the dielectric loss does not follow the percentage of deteriorated oil added. Figure 8 shows this relation. The increase in temperature required to bring the heavier oil to the same viscosity causes thermo-agitation and results in further ionic dissociation in the oil, principally from the impurity content, which accounts for the unusually rapid rise of dielectric loss at the higher percentages of the added deteriorated oil.[7]

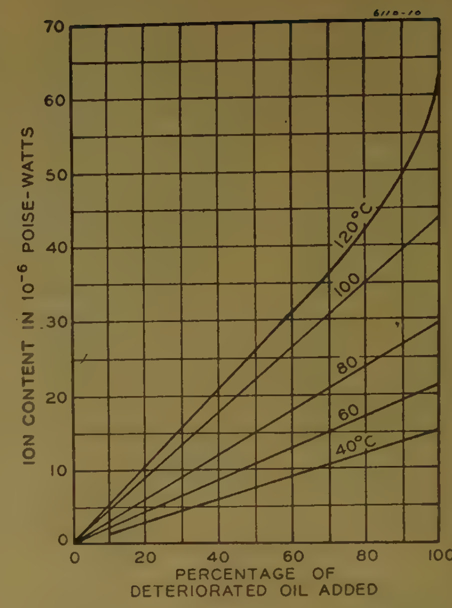

Figure 10. The influence of deteriorated oil on ion content at constant temperature for low viscosity mixtures

In order to substantiate the assumption that temperature increase will increase the concentration of ions in the oil, two sets of curves are plotted in Figure 9, using ion content as a function of temperature for each mixture. The values of ion content for different samples are taken as that which was suggested by Whitehead, namely, the product of conductivity and viscosity.[8] For want of a suitable unit to express ion content, these values are expressed in poise-watts. In these particular cases, dielectric loss is used instead of conductivity. As it has been seen in equation 1 that for a constant value of applied voltage, the dielectric loss is a linear function of conductivity. The curves will have the same slope whether dielectric loss or conductivity is used as the ordinate. If the temperature rise should have no effect on the ion content, these two sets of curves ought to be horizontal lines. In the actual cases the curves are found to rise with temperature, having almost similar constant slopes for each group. Based on the average slope of these curves, the relations may be expressed in the equation

$$\log \frac{N_2}{N_1} = 0.008T \qquad (8)$$

where

N_1 is the ion content at temperature 1
N_2 is the ion content at temperature 2
T is the difference in temperature in degrees Kelvin

In a more general form, this expression may be written as

$N_2 = N_1 e^{cT}$

A New Electronic Timing Relay for Reducing Outages on Power Circuits

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Synopsis: This paper describes a new timing relay for controlling relaying time on automatically reclosed circuit breakers on which inverse-time overcurrent relays are supplemented by instantaneous overcurrent relays. This device incorporates the advantages of instantaneous circuit breaker opening on the first, and in some cases the second, tripout to clear transient faults, and inverse-time opening on subsequent tripouts to obtain co-ordination with branch or sectionalizing fuses. It was designed for universal application on feeder circuits of the Oklahoma Gas and Electric Company system. Certain features which provide for optimum service reliability were incorporated in the design.

EVEN THOUGH all possible protective measures of prevention have been taken in the design of overhead circuits, faults, the majority of which are transitory in nature, still will occur. In order to minimize damage to equipment and interruption to service, the faulted section must be isolated promptly from the rest of the system. Experience has shown that transmission and distribution circuits, on which fault duration is limited to a maximum of 0.2 second, will seldom be out of service for long periods of time as a result of damaged insulators or burned conductors at the point of fault, except in peculiar cases of rare occurrence. Thus, the speed with which protective relays and circuit breakers operate to clear abnormal conditions has a direct bearing on service reliability.

The practice of instantaneous tripping of the source breaker for a fault at any point on the circuit is, by definition, a protective measure of amputation. But, when supplemented by rapid automatic reclosing, its effect on the ultimate load entitles it to be classified as a protective measure of mitigation. It limits the occurrence of permanent faults which are initiated by transient disturbances. In addition, it restricts branch or sectionalizing fuse operation, where possible, to faults of a permanent nature, and thereby reduces long outages on branches or sections of the circuit. Consequently, the realization of the advantages of instantaneous tripping supplemented by immediate initial reclosure has furnished an incentive for designing many ingenuous relaying schemes.

Probably the most effective and certainly the most economical method of securing high-speed relaying on circuit breakers serving multibranch fused circuits is that of superimposing controlled instantaneous overcurrent relays on inverse-time overcurrent relays, and utilizing immediate initial reclosure. This applies to the ground relay as well as the phase relays on grounded neutral circuits. The instantaneous relays are set, where possible, to initiate circuit breaker clearing of faults at any point on the circuit. When these relays are utilized for the second as well as the initial tripout, inrush magnetizing current transients following the initial reclosure may place a minimum limit on the pickup setting.

The co-ordination desired from this type of protective scheme is protection to the entire circuit during the first, and in some cases the second, tripout, in order to give temporary faults, such as flashovers resulting from lightning or momentary conductor contacts, an opportunity to clear before permanent damage occurs and before a branch or section is interrupted by fuse operation. Thus, in the case of faults which are cleared on the initial tripout, a momentary service interruption is substituted for a long outage from damaged equipment or blown fuses. In effect, this momentary service interruption is no more adverse than the voltage

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This gives further substantiation that the ion content is a function of temperature.

Following this indication, it suggests that if temperature should be kept constant, the value of ion content would be a definite function of the percentage of deteriorated oil added. Curves plotted using ion content as a function of percentage of deteriorated oil added at each temperature are shown in Figure 10. These curves show definitely that the ion content is a linear function of the percentage of deteriorated oil added when temperature is reduced to a constant.

Conclusion

This study shows that

1. Provided there is no dipole loss, the relationship between the dielectric loss W and viscosity η of an insulating oil may be expressed as

$$\log W = A - B \log \eta$$

which points out that the mobility of ions is an important factor in dielectric loss of an insulating oil.

2. Thermo-agitation, which causes further dissociation of the oil, increases dielectric loss by increasing the number of ions present.

3. Dielectric loss is increased rapidly by the addition of deteriorated oil. This bears out that dielectric loss is a function of the concentration or content of free ions in the oil.

4. An increase in concentration of ions in an oil may result in an actual decrease in dielectric loss if the viscosity is increased to the point where the mobility of the ions is decreased relatively more than the ion concentration is increased.

5. If nonpolar oils of the same degree of "electrical purity" are used for the low frequency or 60-cycle applications where the flow of the liquid is not important, it is seen that an oil of higher viscosity is preferable from the standpoint of lower dielectric loss.

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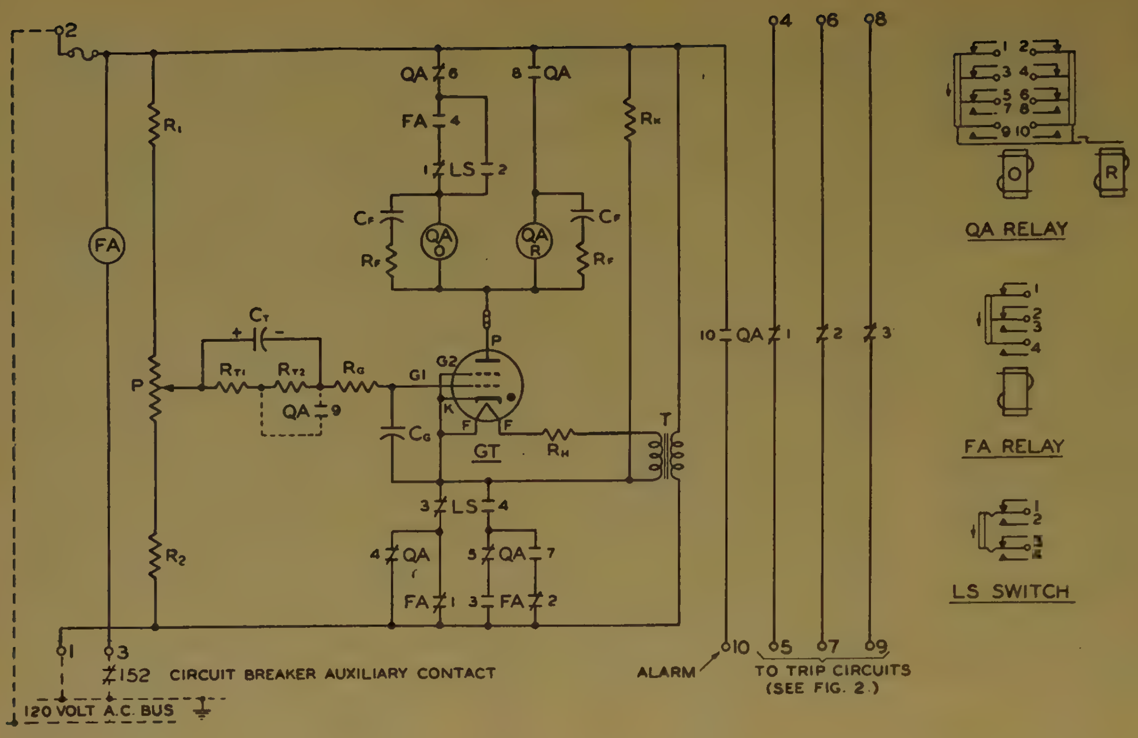


Figure 1. Schematic diagram of the electronic timing relay assembly

disturbance created by a longer clearing time of a fuse. When instantaneous relay protection is also utilized on the second tripout, the time interval between the initial and the second reclosure should be made as short as the imposed circuit breaker duty will permit in order to minimize outage time. Thus, in the case of transient faults which are not cleared initially but disappear on the second tripout, a 30- to 60-second service interruption is substituted for a probable long outage on some portion of the circuit.

To provide backup protection and selectivity between the circuit breaker and the branch or sectionalizing fuses in the event of a permanent fault, the instantaneous relays must be removed from service prior to the initial, or in some cases the second, reclosure, and the inverse-time relays depended upon for subsequent tripping. Through time–magnitude co-ordination, the inverse-time relays will permit branch or sectionalizing fuses to clear the faulted section before initiating breaker operation for backup protection. Where sufficient co-ordination can be realized, these relays are seldom called upon to clear a fault.

Also, in the event of punctured insulators or lightning arresters, the prolonged arcing time secured on the inverse-time relays may produce additional damage and perhaps shatter the faulty equipment. This will permit the fault to be more easily located by patrolmen.

Once removed from service, the instantaneous relays should be held inoperative for a definite time after a reclosure as assurance that the fault was not permanent; otherwise, in the absence of selectivity a lockout of the circuit breaker will occur, resulting in a needless long outage on the entire circuit. To accomplish this, a preset time delay must be interposed between each reclosure and reestablishment of the instantaneous relays. This reset timing is as important in preventing long outages as the instantaneous relays themselves. It should be the minimum time necessary to determine a successful reclosure in order to provide instantaneous relay protection against the contingency of rapidly recurring transient faults.

It is apparent from the foregoing that an auxiliary device capable of the following functions is required:

It must remove the instantaneous relays from service prior to the initial, and in some cases the second, reclosure.

It must hold them inoperative until a successful reclosure is assured and then reinstate them to re-establish initial conditions for future disturbances.

Description of the Timing Relay

In view of the fact that a device fulfilling the foregoing requirements was not available and that existing methods of changing relaying time have certain undesirable features, an electronic timing relay was designed for universal application with relaying schemes utilizing instantaneous and inverse-time overcurrent relays. Briefly, it consists of an assembly of two telephone type relays, a thyratron timing tube, and associated electronic control components mounted in a small drawout case. Figure 1 is a schematic diagram of the assembly in which:

Thyratron *GT* is a gas tetrode, similar to type 2050, operated as a negative control tube. Its operation is discussed under "Theory of the Timing Circuit."

Relay *QA* is a telephone type electric-interlocking relay. It consists of two quick-acting relays mounted as a unit with their armatures mechanically interlocked. When lockup relay *QA-O* is energized, its armature latches and holds its contacts in the operated position until the release relay *QA-R* is energized to trip the latch. The release relay *QA-R* does not latch, but drops out when it is de-energized. Since these relays are pulse operated through their own contacts, a filter consisting of capacitor C_F in series with resistor R_F is shunted across each coil to assure positive operation.

Relay *FA* is a telephone type a-c relay operated as an equivalent circuit breaker auxiliary switch. Its unoperated position corresponds to the closed position of the circuit breaker.

Lever Switch *LS* is a double-pole double-throw selector switch which provides the selection of either one or two instantaneous relay operations.

Transformer *T* is a voltage dropping transformer for supplying the heater of the thyratron.

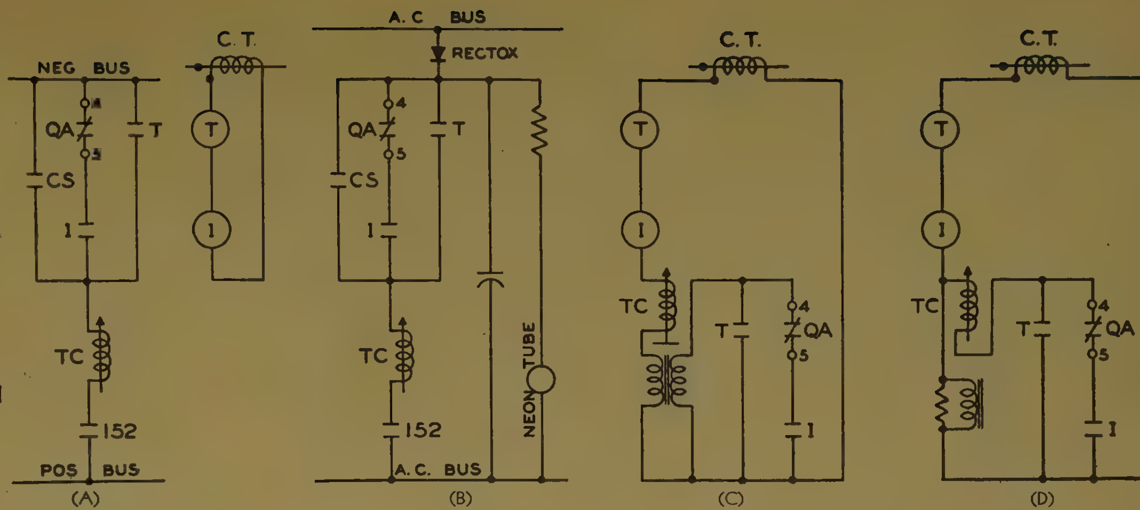
Other components: Resistors R_H and R_G are merely current limiting resistors. Capacitor C_G provides protection against high voltage transients. The functions of the other components are discussed with reference to Figure 3 under "Theory of the Timing Circuit."

To avoid confusion, the graphical symbols used in all the diagrams are, insofar

Figure 2. Location of electronic timing relay contacts in breaker tripping circuits

I—Instantaneous overcurrent relay
T—Inverse-time overcurrent relay
QA—Electronic timing relay (Figure 1)

A—D-c method
B—Capacitor method
C—Transformer method
D—Reactor method



as possible, those recommended by American Standards Association. Device contacts are shown for the de-energized or unoperated position of the device. All tube voltages are referred to the equipotential cathode. Instantaneous values of varying electrical quantities are represented by lower case letters, and their peak values by capital letters. Where possible, subscripts are chosen so as to be indicative of that portion of the circuit to which they apply.

Theory of the Timing Circuit

The basis of the timing relay design is the electronic timing circuit shown schematically in Figure 3, in which timing action is obtained by grid amplitude bias of the gas tetrode *GT*, acting as an inertia-less circuit closing or opening device in series with relay *QA*.

An alternating voltage e_L of commercial frequency is impressed between *L*1 and *L*2, and the tube heater circuit is energized from a filament transformer (not shown). The fixed resistors R_1 and R_2 in series with potentiometer *P* form a voltage divider circuit for controlling the potential applied to the control grid *G*1. With switch *S*1 open and switch *S*2 closed, unidirectional pulsating electron current flows from the cathode *K* to the grid *G*1 during the half cycles in which *L*1 and the grid are positive with respect to *L*2 and the cathode. As a result of this rectification, a charge of the indicated polarity will accumulate in the timing capacitor C_T and which will be maintained as long as switch *S*1 is open. Also, with switch *S*1 open, the cathode *K* is connected through resistor R_K to line *L*2 and is approximately at the same potential as the anode *P*; consequently, no electron current flows from cathode to the anode to energize the relay even though switch *S*2 is closed.

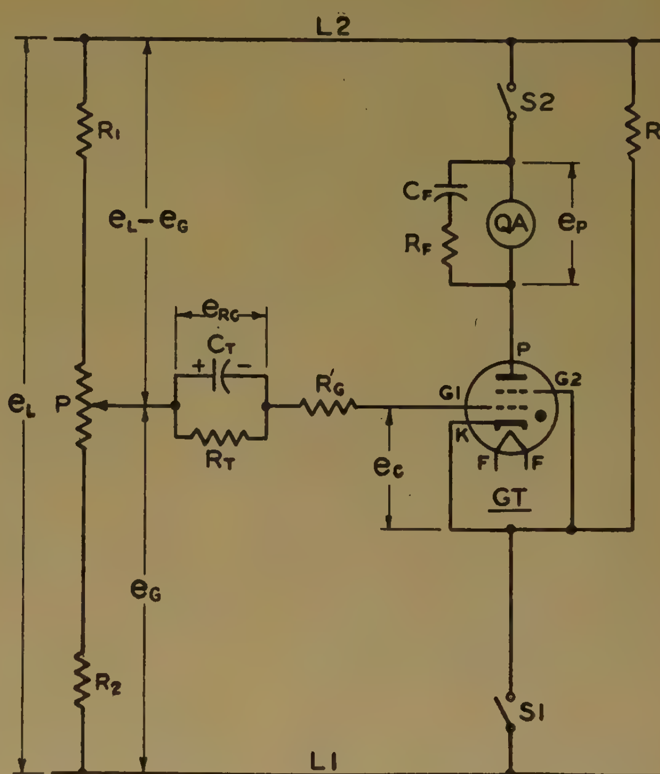


Figure 3. Schematic diagram of the electronic timing circuit

When switch *S*1 is closed, the cathode is connected directly to line *L*1 which impresses voltage e_L across the anode–cathode circuit and superimposes an alternating voltage e_G (in phase with the anode voltage) on the negative grid bias voltage E_{RC} of the timing capacitor C_T. The alternating component e_G of the grid bias voltage e_c tends to drive the grid positive at the same time the anode is positive; however, the tube does not conduct between cathode and anode until the grid bias voltage e_c is equal to or more positive than the critical ionization value determined from the tube's control characteristic.

The instantaneous relationship of the anode and control-grid voltages is shown graphically in Figure 4, in which time is reckoned from the instant switch *S*1 is closed, and the alternating voltages are zero and increasing in the positive direction. It is apparent from this graphical representation that the magnitude of the grid bias voltage is a time function of both the unidirectional component e_{RC} and the alternating component e_G and defined by the equation

$$e_C = e_{RC} + e_G \tag{1}$$

in which

$$e_{RC} = E_{RC}\epsilon^{\frac{-t}{R_T C_T}} \tag{2}$$

is the solution of the differential equation defining the time-voltage characteristic of

capacitor C_T discharging through resistor R_T, where

E_{RC} = potential difference across C_T at the time discharge is initiated ($t=0$) and is approximately equal to $E_L - E_G$ if $R_G + R_K$ is small in comparison with R_T

C_T = capacitance in farads

R_T = resistance in ohms

t = time in seconds from initiation of discharge

ϵ = base of natural logarithms (2.718 . . .)

Solving equation 2 for time gives

$$t = R_T C_T \log_\epsilon \frac{E_{RC}}{e_{RC}} \qquad (3)$$

from which the time required for E_{RC} to decay to a desired value of e_{RC} may be calculated.

Since the values of the timing capacitor C_T and discharging resistor R_T are fixed, the time constant is fixed and the time range is determined by the setting of the potentiometer P and the relative values of the components in the voltage divider circuit. The upper end of the potentiometer determines the minimum peak voltage to which the timing capacitor charges before initiation, the maximum peak voltage of the grid bias component after initiation, and therefore the minimum time delay between initiation and energization of relay. Conversely, the lower end of the potentiometer determines the maximum peak voltage to which the timing capacitor charges before initiation, the minimum peak voltage of the grid bias component after initiation, and thus the maximum time delay between initiation and energization of relay.

After expiration of a preset time delay, the tube "fires" and conducts current in the anode–cathode circuit during the half cycles in which the anode is positive with respect to the cathode, energizing relay QA and charging the filter capacitor C_F. The relay is held in during the nonconducting half cycles by the energy stored in the filter capacitor. Thus, the relay will remain picked up until the anode–cathode circuit is opened by switch $S1$. Opening switch $S1$ re-establishes initial time ($t=0$) conditions. By properly co-ordinating the circuit components, the reset time requires only a few cycles, after which the timing cycle may be repeated.

By merely reversing the sequence in which switches $S1$ and $S2$ are operated, essentially instantaneous energization of relay QA is obtained. With switch $S2$ open and switch $S1$ closed, the peak voltage to which timing capacitor C_T charges is approximately equal to the peak voltage of the alternating grid bias component, so that the grid is approximately at cathode potential during the positive peaks of the alternating grid bias component. That is, E_{RC} is approximately equal and in phase opposition to E_G; consequently, after switch $S2$ is closed, ionization will occur on or near the first positive peak of e_G.

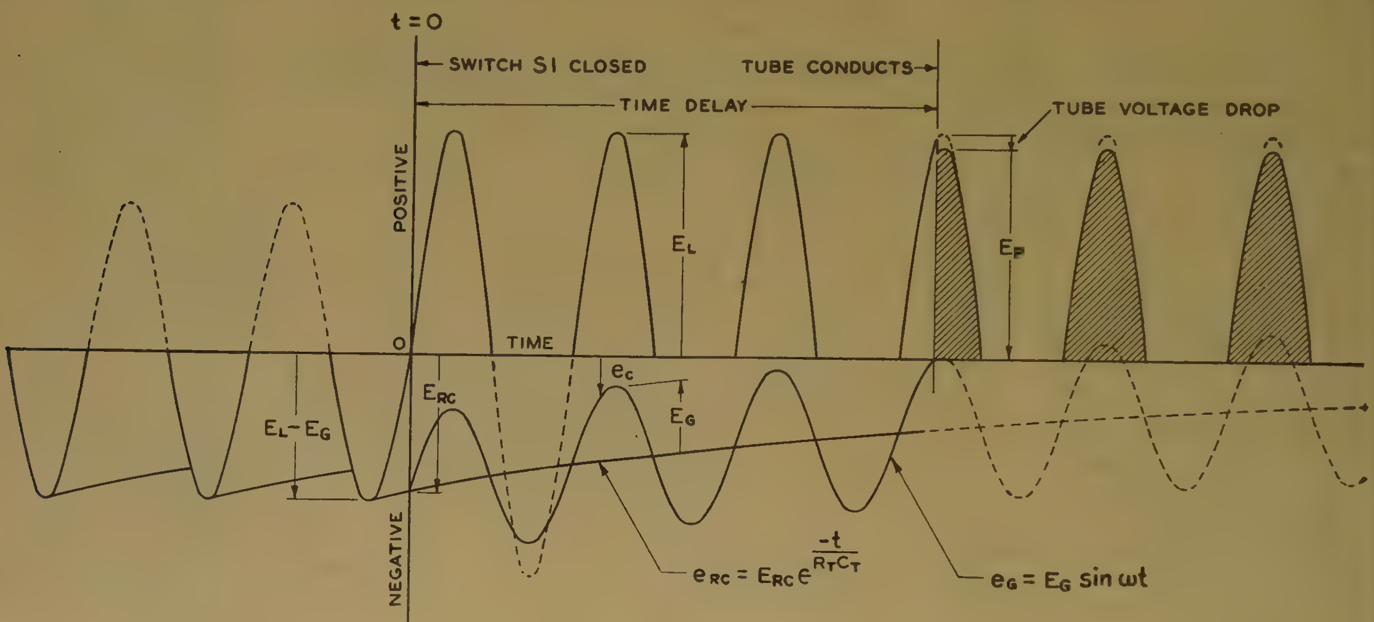


Figure 4. Instantaneous relationship of anode and control grid voltages

External Connections

The timing relay is operated from any 120-volt a-c source connected across terminals 1 and 2, and it is controlled through a contact of the circuit breaker auxiliary switch (152) connected between terminals 1 and 3. These connections are indicated on Figure 1 by the broken lines. In a majority of cases the timing relay may be controlled directly from the circuit breaker indicating light or alarm circuits since the only requirement is energization of relay FA when the circuit breaker is open.

Contacts QA_1, QA_2, and QA_3 brought out on terminal 4-5, 6-7, and 8-9 are connected into the circuit breaker tripping circuit as shown in Figure 2. Only one of these contacts is required for multiphase potential or shunt trip methods (A and B of Figure 2), since the contacts of all the instantaneous overcurrent relays are paralleled and actuate a single trip coil. But, in the case of multiphase current or series trip methods (C and D of Figure 2) a separate contact for each instantaneous overcurrent relay is required, since these actuate separate tripping elements.

Contact QA_{10} brought out on terminal 10 may be used for initiating an alarm or operation counter.

Operation for One Instantaneous Tripout

Where instantaneous relay protection on the initial tripout with inverse-time relay protection on subsequent tripouts is desired, lever switch LS is set in the unoperated "off" position as shown in Figure 1, in which position the timing relay is set up for immediate operation to open the trip circuits of the instantaneous relays prior to the immediate initial circuit breaker reclosure.

When the circuit breaker opens, auxiliary switch contact 152 closes and ener-

gizes relay FA which closes the anode circuit through contact FA_4, and tube conducts immediately to energize lockup relay QA-O. As a result: The trip circuits of the instantaneous overcurrent relays are opened by contacts QA_1, QA_2, or QA_3; lockup relay QA-O is de-energized (but not released) by contact QA_6; release relay QA-R is connected into the anode circuit by QA_8; and the cathode is opened by contact QA_4, allowing timing capacitor C_T to charge for reset timing. All of which occurs within 0.05 second after the circuit breaker trips. This operating time provides definite assurance that the instantaneous overcurrent relays are inoperative in advance of immediate circuit breaker reclosure, since the average operating time from initiation of tripping to reclosure of circuit breakers, such as employed for feeder service, ranges from about 0.5 to 1 second.

On reclosure of the circuit breaker, auxiliary switch contact 152 opens and de-energizes relay FA. This closes the cathode circuit through contact FA_1, and the tube conducts after a preset time-delay to energize release relay QA-R which releases lockup relay QA-O. As a consequence, the timing relay will reset to re-establish initial conditions if the circuit breaker remains closed through a timing cycle. However, in the case of faults which are not cleared on the initial tripout, the circuit breaker will be tripped by the inverse-time overcurrent relays, and the timing relay will remain in its operated position to maintain inverse-time tripping until the circuit breaker is successfully reclosed. The reset time delay should be the minimum necessary to co-ordinate with the circuit breaker clearing time on the inverse-time overcurrent relays, in order to provide instantaneous relay protection against the contingency of rapidly recurring faults.

Operation for Two Instantaneous Tripouts

Where instantaneous relay protection on the second, as well as the first, tripout is desired, lever switch LS is set in the "on" position, so that contacts LS_1 and LS_3 are open, and LS_2 and LS_4 are closed. In this position the timing relay is set up for time-delay operation, since when the cathode circuit is open through contacts FA_3 and QA_7, timing capacitor C_T is precharged for timing action.

When the circuit breaker opens, auxiliary switch contact 152 closes and energizes relay FA which closes the cathode circuit, but the tube does not conduct until after the preset time delay. If the immediate initial reclosure is successful, relay FA is de-energized, and initial conditions are re-established. However, if the fault is not cleared initially, this time delay allows the circuit breaker to be tripped again by the instantaneous overcurrent relays.

In the case of faults which are not cleared on the second tripout, the tube "times out" and then energizes lockup relay QA-O. As a result: The trip circuits of the instantaneous overcurrent relays are opened by contacts QA_1, QA_2, or QA_3; lockup relay QA-O is de-energized (but not released) by contact QA_6; release relay QA-R is connected into the anode circuit by contact QA_8; and the cathode circuit is opened by contact QA_5 which permits timing capacitor C_T to charge for reset timing. All of which occurs prior to the second reclosure, so that subsequent tripping of the circuit breaker is initiated by the inverse-time relays.

Following a successful reclosure, the timing relay will reset in a manner similar to that described under "Operation for One Instantaneous Tripout."

Since the timing adjustment (potentiometer P) is common to both operation and reset, it must be such that the preceding requirements are fulfilled. In some cases, it may be desirable to have less time-delay on reset than on operation, and this is accomplished by short-circuiting a section of the timing capacitor discharging resistor as shown by the dotted connection of contact QA_{10} across resistor R_{T2} in Figure 1. The reverse can be accomplished by making QA_{10} a normally closed contact.

Conclusions

The most salient features of the timing relay are:

1. Universal application regardless of circuit breaker mechanism or method of control.

2. Selection of either one or two instantaneous tripouts.

3. Independent of the control source voltage during fault conditions.

4. Reset timing independent of reclosing relay.

5. Ease of installation.

The initial installation of this equipment was made February 1945 on a breaker serving a high exposure 24-kv grounded Y circuit. As a result long time outages previously experienced on this circuit practically have been eliminated.

Sixty similar installations will be made on the Oklahoma Gas and Electric Company system during 1946.

With certain changes in the internal wiring, this device can be converted into an instantaneous or time-delay "one-shot" recloser with quick reset on a successful reclosure.

Silicone-Resin-Treated Magnet Coils

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MAGNET COILS present somewhat different insulation problems from rotating machines. This was recognized when separate magnet coil standards[1–3] were set up by AIEE, National Electrical Manufacturers Association, and American Standards Association which permit higher maximum observable temperatures than do the rotating machine standards. AIEE Standard 1 states, "How long an insulation will last will depend not only on the class of material used but also upon the physical support of the insulation and the severity of the physical forces tending to disrupt it." Following this line of reasoning, AIEE Standards 15 and 16 permit higher observable temperatures on magnet coils as shown in Table I. Work by Erikson[4] indicates the desirability of separate consideration of magnet coils and suggests the possibility of higher operating temperatures even with organic insulation.

This distinction became more pronounced with the development of silicone insulation. Magnet coils were therefore considered as a special study of the basic work previously reported[5,6] on the general application of silicone varnishes to electric machinery. The test on coils was planned with the view of determining:

(*a*). The temperature–life relationships of the silicone varnish.

(*b*). The maximum operating temperatures of magnet coils for normal life expectancy.

(*c*). The effect of aging at various temperatures on the thermal conductivity of the assembly.

(*d*). Design advantages offered by higher temperature operation.

Test Program

Paralleling the motor test program[7] a like series of tests was planned for magnet coils. Briefly this program consisted of operating coils at different temperatures to determine the thermal life expectancy of the coils. In order that these tests might be useful and representative of standard practice, a commercial contactor and coil design were chosen. This is a conventional magnetic contactor for d-c control circuits which is widely used for industrial and railway service. The general construction is shown in Figure 1. All tests were run on this design of contactor in order that results obtained would be subject to direct comparison.

The operating temperatures from 200 to 275 degrees centigrade were chosen. The top limit was chosen because it was anticipated that at 275 degrees centigrade failures by short-circuited turns would probably occur in a few weeks. Another factor limiting the test to 275 degrees centigrade was the melting point of the solder (304 degrees centigrade) which held the coil leads to the terminals. Early estimates based on extrapolated curves for silicones indicated that 200 degrees centigrade would be the maximum operating temperature resulting in the full life expectancy of the coils. Thus the tests were started with 12 coils, three each at 200, 225, 250, and 275 degrees centigrade.

The data desired were to be obtained by day to day operating temperature readings (by resistance), mechanical stressing by thermocycling and operating the contactors, periodic measurement of cold resistances to determine wire oxidation and possible short-circuited turns (indications of failure), notation of color changes of resin on surface, and power input.

Coil Design

The coils tested were essentially a standard mechanical design to simulate normal conditions. They were wound with number 25 (0.0179 inch diameter) single glass covered copper wire (silicone varnish treated). Silicone wire enamel was not used because at that time the enamel development had not progressed far enough. All winding and insulating details were fiber glass and fiber-glass-backed mica (all bonded and treated with silicone resins). The spool was fabricated of fiber glass laminated tubes with a special resin bond (not silicone) and asbestos phenolic end washers. At the time these coils were made, thermosetting silicone moulding resins had not been developed to a usable state. The terminals were soldered with a high temperature solder (304 degrees centigrade solidifying temperature). No organic materials such as paper, cotton tape, string, glue, shellac, or varnish were used in the coil construction. The complete coil was treated by vacuum-pressure impregnation with DC-990-A silicone varnish, a product of Dow Corning Corporation, at 70 per cent solids and baked for 24 hours at 225 degrees centigrade. This was followed by a dip in 50 per cent solids DC-990-A silicone varnish and baked at 225 degrees centigrade for 16 hours.

The following design data give significant details of the coils used for this test:

Spool—3 1/4 inches long by 1 5/32 inches inside diameter by 3 inches outside diameter.

Washers—1/8 inch thick.

Double insulating tube construction, each 1/16 inch thick.

Magnetic core—1 1/8 inch diameter.

3,500 turns of number 25 wire having a resistance of 61 ohms (25 degrees centigrade).

Winding volume—11.6 cubic inches.

Barrel surface—25 square inches.

Test Procedure

The coils were connected to a 120-volt d-c motor generator set through variable nichrome resistors. Thus the current in each coil could be controlled without affecting the others.

The general testing procedure is detailed in Appendix I. Volt-ampere readings were taken every day, after which the coils were switched on and off ten times. Approximately once every week the cooling curves were run to check the hot volt-ampere resistance readings of each coil. The coils were then left off overnight and then cold resistance readings made the following morning, and the coils again energized.

Changes were made periodically to correct the test conditions for increases in

Table I. Comparison of Permissible Temperature Rise and Observable Temperature for Magnet Coils and for General Purposes

Class	Maximum Permissible Temperature Rise by Resistance, Deg C: General[1]	Magnet Coils[2,3]	Equivalent Maximum Observable Temperature,* Deg C: General[1]	Magnet Coils[2,3]
A	60	85	100	125
B	80	105	120	145

* Rise plus 40 degrees centigrade as maximum permissible ambient.

Paper 46-104, recommended by the AIEE committee on industrial control devices for presentation at the AIEE summer convention, Detroit, Mich., June 24–28, 1946. Manuscript submitted November 13, 1945, made available for printing April 11, 1946.

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Figure 1. Laboratory test setup showing magnet coils under test being examined by Mr. Torok

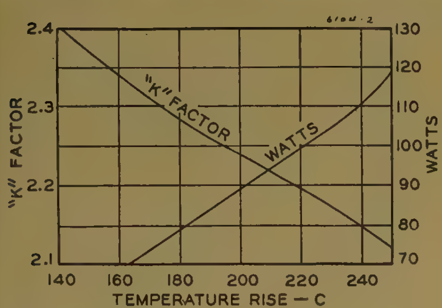

Figure 2. Effect of temperature rise on K factor and watts

cold resistance of the winding due to oxidation of the copper. After one year of operation without a failure, it was believed that the test could be made more severe by additional thermal cycling of the coils. The thermal cycling consisted of switching the power off the coils twice each day. The coils would be de-energized for approximately two hours in which time the coils would cool to a little above room temperature. Then power would be applied for four hours, a period more than sufficient for them to reach stabilized operating temperature. The power was then turned off for two hours and then left on overnight. Once each week the power would be off overnight and a cold temperature reading made the following morning.

Test Results

The change in heat dissipation factor K was used as a basis for observing insulation aging. This K factor is the average temperature rise divided by watts input and represents the temperature the coil must attain to dissipate the energy liberated within its body. Since average temperatures by resistance were used, the heat dissipating ability of the boundaries of the coil was only one factor involved. The heat transmission within the coil has an appreciable influence as it determines the distribution of temperature within the body of the coil. If the heat conductivity of the body is low, the internal temperature may be much higher than the surface temperature, where the heat is dissipated. Thus the average temperature being higher, the coil would have a very high K factor. However, if the heat conductivity within the coil were very good, the surface and internal temperature would be more nearly the same and a low K factor would result. Thus a change in K factor may represent a change in the rate of heat dissipation, but more likely the change in thermal conductivity within the body of the coil.

Aging and heat deterioration in organic and semiorganic materials generally are accompanied by weight loss, which results in dimensional changes or the formation of voids. The resinous bonds between the turns in the coil act not only as physical supports but as thermal conductors. As the impregnant ages, it shrinks or otherwise forms voids and its thermal conductivity diminishes. Resins shrink slightly as they cure and cause some voids which manifest themselves as a rise in K factor. Figure 2 shows the original K factor as a function of temperature rise above 25 degrees centigrade ambient temperature. Figure 3 shows the per cent increase in K factor as a function of hours aging under each temperature condition.

After the tests had been in progress, changes in cold resistance were observed. Thereafter periodic observations of cold resistance were made and the changes recorded. Figure 4 shows the per cent change in cold (25 degrees centigrade) resistance as a function of time of aging at various temperatures.

No definite coil failures occurred. Only two coils encountered trouble during the whole test and these were the result of faulty soldered joints at the terminal. They were detected by variable resistance readings.

Figure 5 shows the conditions of coils 2, 4, 5, 8, 9, 10, 11, and 12 after the conclusion of the test. All the coils except number 4 had been at temperature 12,615

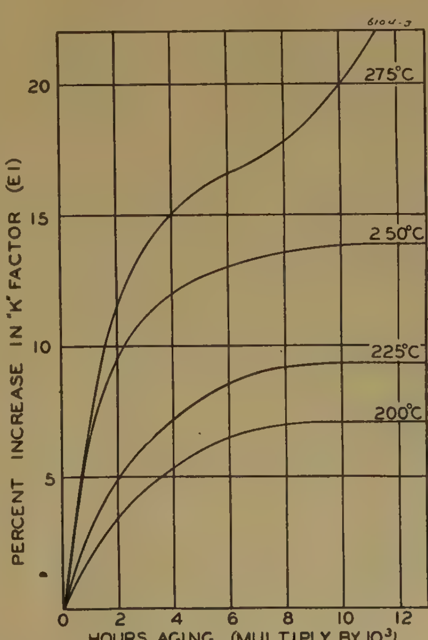

Figure 3. Effect on K factor of aging at elevated temperatures

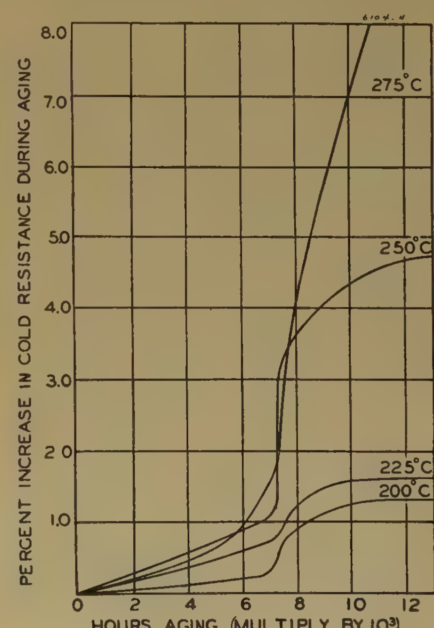

Figure 4. Effect of aging at elevated temperatures on the cold (25 degrees centigrade) resistance of magnet coils

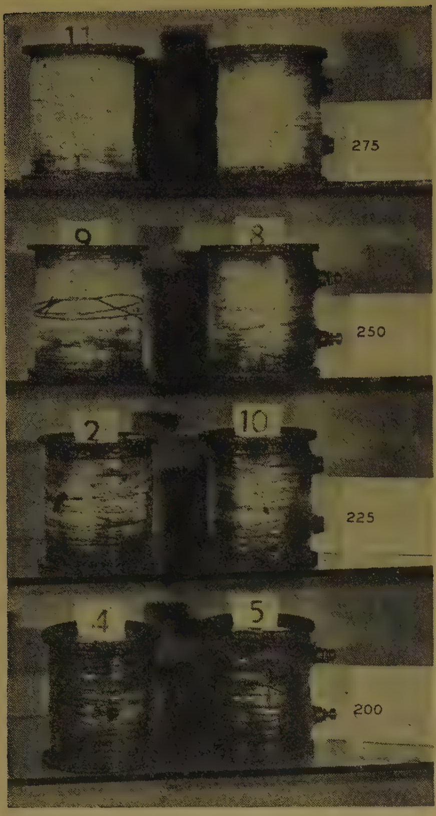

Figure 5. Coils showing external condition after 12,000 hours aging at temperatures of 200, 225, 250, and 275 degrees centigrade

hours. Coil 4 had been at temperature 11,006 hours.

The coils shown in Figure 5 were subjected to electrical tests as follows:

(a). All passed the standard low-voltage short-circuit test ("E" iron balanced watt meter tester).

(b). All the coils except numbers 11 and 12 successfully withstood surge comparison tests[8] up to 10.6 kv (the limit of the tester available). Coils 11 and 12, which were aged 12,615 hours at 275 degrees centigrade, failed at 8 kv. These test voltages were surge peak values across two coils in series.

After electrical tests, the coils were cut open for physical examination. The condition of typical coils is shown in Figure 6. Detailed observations are given in Table II. These may be summarized in the following statements:

(a). Coils at 200 degrees centigrade were in excellent condition showing few if any signs of aging.

(b). Coils at 225 degrees centigrade were beginning to age, but were well bonded and in good condition.

(c). Coils at 250 degrees centigrade showed unmistakable signs of aging, but were still in fair condition, being well bonded but brittle.

(d). Coils at 275 degrees centigrade were still operable. However, they were in a condition which might be expected to result

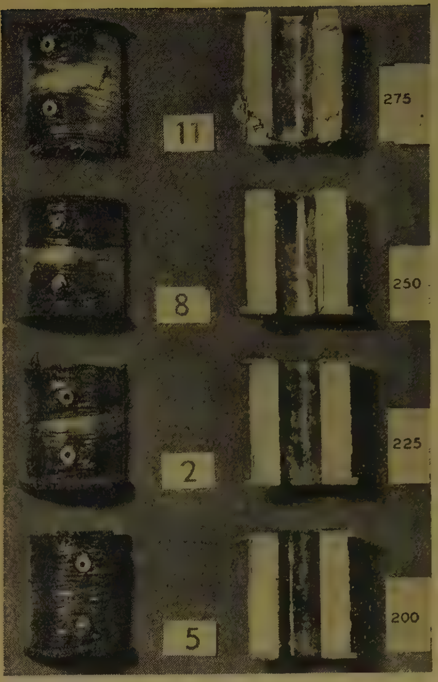

Figure 6. Coils showing internal condition after 12,000 hours aging at temperatures of 200, 225, 250, and 275 degrees centigrade

in failure under actual service conditions, as the bonding between wires as well as parts of the spool was near the failure point

Thermal decomposition of silicones produces a reversal of color change from that normally encountered in organic resins. Organic resins decompose to carbon, a black conducting substance, whereas silicones decompose to silica, a white insulating substance. The aging of these coils clearly shows this color change which was an important indication. (The real significance of this color change requires further study.)

Examination of Copper Conductor

One of the coils which had operated on this test for 9,000 hours at 275 degrees centigrade was cut open for examination. The microstructure of this wire was normal for electrolytic tough-pitch copper and no hydrogen embrittlement could be detected. A black coating was present on the wire (about 0.0004 inch thick) which was found to be a mixture of cuprous and cupric oxide by X-ray diffraction analysis. The X-ray pattern for this copper was normal.

It was concluded that the test conditions had no deleterious effect on the copper wire except for slight oxidation of the surface. This surface oxidation resulted in increase in conductor resistance.

Discussion of Test Data

Over 12,000 hours ($1^1/_3$ years) operation at temperatures of as much as 275 degrees centigrade did not destroy a single coil or produce a case of serious trouble (except the two which were attributable to original workmanship in soldering).

The change in heat dissipation factor K shows an upward trend as the coils age. The rate of increase in K factor is greater at higher temperatures as shown in Figure 3. The initial rise in K factor may be considered as indicating completion of internal cure. However, the abrupt break in the curve for coils at 275 degrees centigrade (at 8,000 hours) is believed to indicate thermal degradation.

Resistance changes, though definite, were not of a serious nature even on coils at 250 degrees centigrade for the full period of the test. At 275 degrees centigrade the resistance change shown in Figure 4 is approaching a serious condition indicating that the temperature limitation may be established by copper rather than insulation. The rate of copper oxidation is known to increase as the temperature rises. Since copper oxide at low potentials is effectively an insulator, the rate and extent of oxidation can be determined by the change in resistivity of the copper wire. Periodic coil resistance readings were necessary to determine the extent of oxidation of the windings. The cold resistances remained essentially unchanged until the test had progressed to above 6,000 hours, then they started to rise and approach a constant level except at 275 degrees centigrade. At that temperature, no saturation level was observed. This level rose at an increasing rate as the temperature

Table II. Physical Condition of Magnet Coils After Thermal Aging for Approximately 12,000 Hours

Aging Temperature, Deg C	Coil Numbers	Condition of Washers	Condition of Winding: Color	Condition of Winding: Bond Strength
200	4–5	Well bonded	Dark	Excellent
225	2–10	Well bonded	Beginning to lighten	Good
250	8–9	Fairly well bonded	Light	Still bonded but brittle
275	11–12	Loose	Light	Poor—ready to fall apart

rose above 200 degrees centigrade. This seems to indicate that the copper oxide forms a protective film over the copper which decelerates the oxidation for a given temperature. That the oxidation did not appear until about 6,000 hours as compared to an almost immediate rise in K factor also indicates that the silicones' adhesion to copper is good and it is the last place where voids begin to form. The extent of oxidation might be greatly reduced if the bare copper wire had a protective coating, such as cadmium plating, under the insulation.

Interpretation of Data

One of the main objectives of this program was to obtain some idea of the order of magnitude of the thermal life of the silicone resins and composite silicone insulation. A recent review[11] of the aging characteristics of insulation disclosed general agreement that thermal aging of insulation is a logarithmic function of temperature. This concept was utilized in extrapolating the results of this series of tests. Extrapolation of short time tests is recognized as difficult but if predictions are to be made, it offers at least an opportunity for an "intelligent guess."

Figure 7 has been constructed as follows to extrapolate these data:

(*a*). The actual test hours at the various temperatures were plotted as points A, B, C, and D.

(*b*). A suggested life curve for silicone insulated motors was plotted for reference, with a 12 degree centigrade slope for half-life. (By coincidence this passes through point B.)

(*c*). From examination and tests, it was concluded that there is little likelihood of failure at the condition resulting from the 250 degrees centigrade test (point C). However, the condition resulting from the 275 degree centigrade test (point D) was interpreted to be near the end point. These points were therefore assumed to represent minimum and maximum life and were used to locate the boundaries of the "probable failure zone." Curves then were drawn through these points parallel to the suggested life curve for silicone insulation on motors.[7]

(*d*). A reference bench mark is drawn at 7 years and reference points are shown on this line for class B motors and class B magnet coils, at temperature limits established by AIEE Standards, as indicating probable life expectancy.

From this curve, it can be seen that the probable "minimum life" curve for magnet coils intersect the bench mark line at 225 degrees centigrade. It is of interest that this is 25 degrees centigrade above the probable motor insulation life curve which is the established difference for magnet coils over motors for class A and class B insulation (see Table I). Thus the test results indicate that the increase in temperature for comparable life is consistent between motors and magnet coils with silicone insulation when compared to the present allowed temperatures for class B motors and magnet coils.

The proposed temperature allowances[6] for silicone motor insulation corresponding to AIEE Standard 1 are 160 degrees centigrade observable by resistance. This is well below the insulation life value indicated by tests on motors.[7] It, therefore, is consistent and conservative to suggest a permissible temperature for silicone magnet coil insulation below the indicated value of 225 degrees centigrade by resistance. It is the opinion of the authors that a value of 200 degrees centigrade as observed by resistance would be satisfactory.

Effect of Operating Temperature on Design

The permissible operating temperature is an important factor in magnet coil design. Temperature rise above ambient temperature can be approximated at conventional operating temperatures for a given coil design by the following formula:

$$E=\sqrt{\frac{R\times\Delta t}{K}\left(1+\frac{\Delta t}{234.5+t_a}\right)}$$

where

E = continuous rated voltage for temperature rise of Δt

R = cold resistance of winding at temperature of t_a

Δt = temperature rise by resistance above t_a

t_a = ambient temperature (for operating or test conditions)

K = heat dissipation constant of coil design, in degrees rise per watt

This formula assumes that the temperature rise is proportional to the watts generated in the coil winding. This is reasonably accurate until the higher temperatures are reached when the K factor tends to decrease (heat dissipation is not a linear function of temperature rise).

The relation between continuous voltage rating and temperature rise is shown in Figure 8 based on the foregoing formula. This curve also shows that the ampere turns, while increasing with rated voltage, do not increase in the same ratio because of the increase in winding resistance at the higher temperature.

From Figure 8, it is clear that the present difference between class A coils at 85 degrees centigrade rise, and class B coils at 105 degrees centigrade rise (assuming the same space factor), is somewhat less than nine per cent improvement in ampere turns. This difference is not of sufficient importance to justify the use of class B magnet coils on most applications, hence they are not used generally. To obtain a significant advantage from increase in operating temperature (with a standard size of coil), it would be necessary to increase the permissible rise from 85 degrees centigrade to approximately 160 degrees centigrade (about equivalent to one standard wire size).

Figures 9 and 10 illustrate the advantage to be gained in reduced size of winding through increase in temperature rise.

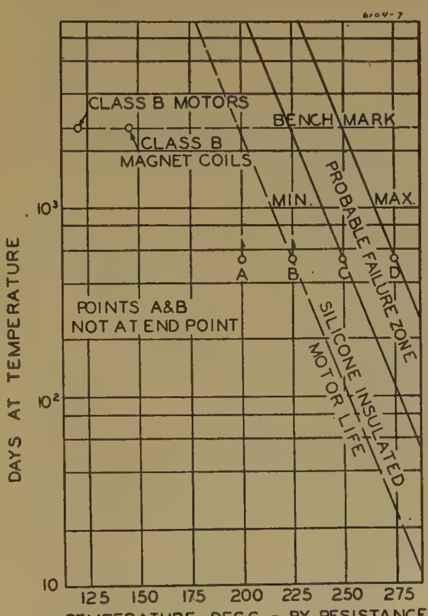

Figure 7. Thermal aging curve interpreting test results

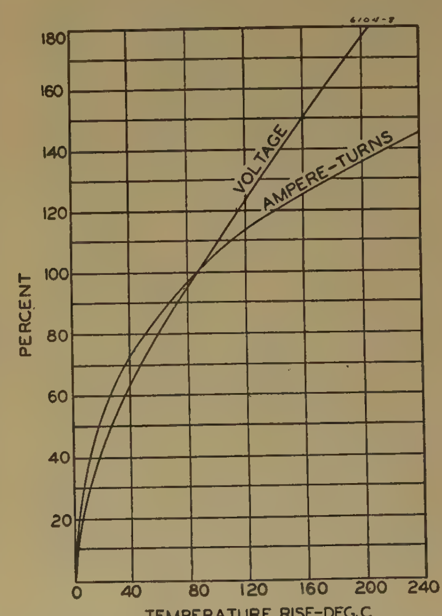

Figure 8. Effect of temperature rise on voltage and hot ampere-turns of magnet coils

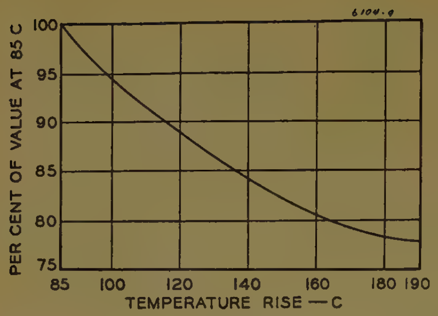

Figure 9. Effect of temperature rise on active coil length and copper weight

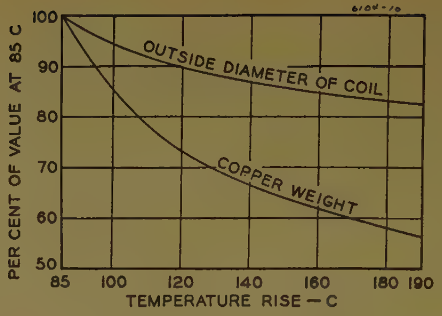

Figure 10. Effect of temperature rise on coil diameter and copper weight

These calculations are based on a constant value of minimum hot ampere turns at a specific percentage of rated voltage. The coils used as the basis for these calculations are the same size and construction as used for the tests reported herein.

Conclusions and Recommendations

One unmistakable conclusion drawn is that the combination of fiber glass and silicone resin makes an important contribution to improving the thermal life of magnet coil insulation. The end of the reliable life of such coils is believed to be determined by the failure of the silicone resins as bonds. The glass fibers provide positive turn separation and the by-products of silicone decomposition are nonconducting. Therefore, the probability of short-circuited turns is not great, even at the condition described herein as "maximum life," unless there is severe mechanical distortion of the winding.

The tests were begun about $2^1/_2$ years ago and employed DC-990-A resin which was the best silicone resin available at that time. Silicone resins have been improved (the corresponding improved resin is known as DC-993). Other short time tests indicate that similar coils with DC-993 will have even greater thermal endurance than indicated by the tests reported herein. Parallel tests employing the improved resins are in progress.

Though the experience is meager, there is considerable evidence to indicate that silicone insulated magnet coils employing fiber glass insulated wires can be rated at temperatures appreciably above the 160 degree centigrade temperature by resistance proposed for silicone insulation on rotating machinery.

Increase in permissible temperature rise will provide a real design advantage through an appreciable increase in net ampere turns. This advantage is obtained at some sacrifice in efficiency, as the watts per ampere turn are increased at elevated temperatures.

It is recommended that other investigators study this problem in order that sufficient confirming data be obtained to warrant establishing a new standard for silicone magnet coils, and particularly that they explore the elevated temperatures where designers may obtain an appreciable advantage (one standard wire size change).

It is the recommendation of the authors that an observable temperature limit of 200 degrees centigrade by resistance be considered as a future standard for silicone insulated magnet coils employing fiber glass conductor insulation. This would permit a temperature rise of 160 degrees centigrade (above 40 degrees centigrade ambient temperature) as observed by resistance.

Appendix I. Test Specification for Temperature Runs on Magnet Coils

Set up complete apparatus with coil assembled in its normal operating position. Shield apparatus from abnormal drafts where necessary but leave unrestricted room for flow of normal ventilating air.

Apply several thermocouples to coil surface. These should be distributed around the surface so as to obtain several readings for an average indication. Make sure that these are in intimate contact with coil surface and that the sensitive element is covered with a small felt pad or insulating cement.

Place a thermometer in room air adjacent to the apparatus to record ambient temperature. This should be near enough to the apparatus to indicate accurately the surrounding room air temperature but should not be in a position where the apparatus temperature will affect it.

Next accurately measure the coil resistance and record it with the coil surface temperature as well as the ambient temperature.

Only after these preliminaries should power be applied. The voltage specified for test should be held constant throughout the test. During the early part of the test, frequent readings should be taken, but in the latter part, as the coil temperature approaches a steady state, less frequent readings will suffice. Heating data should be observed and recorded as follows:

1. Time of reading with reference to cold start.
2. Volts applied to coil terminals.
3. Amperes flowing through coil.
4. Surface temperatures.
5. Room air temperature.

After all observed values reach a steady state (or the increase does not exceed the variation due to change in ambient temperature) take a set of shutdown readings. Arrangements should be made before power is shut off to change the coil connections quickly to a Wheatstone bridge. The time that power is shut off must be observed accurately and related to the cooling data taken. After power is shut off, frequent readings of coil resistance should be made until the change in coil resistance is at least half the total increase due to temperature rise. The coil resistance at the instant of shutdown is determined by plotting the values observed during the cooling period and extrapolating to the time that power was removed.

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Inductive Co-ordination Aspects of Rectifier Installations

AN AIEE COMMITTEE REPORT

THIS REPORT discusses the inductive co-ordination aspects of rectifier installations from the standpoint of effects on the a-c side and presents methods of classifying proposed installations into one of the three following categories:

Category I. Very *unlikely* to cause important induction problems.

Category II. Indeterminate.

Category III. Very *likely* to cause important induction problems.

Suggested dividing lines between categories are set up in terms of estimated harmonic currents or voltages and types of power and telephone systems involved. In using these figures, it is important to keep in mind that the phenomena involved are complex and that simple estimates are necessarily approximate: a difference of less than about 25 per cent in any of the results obtained by these estimates is ordinarily not significant. It is particularly important that none of the figures be construed to be any sort of limit. (For working material, see pages 420 and 421.)

Among the advantages of classification of installations in the manner described herein are:

1. Considerable time and effort can be saved by avoiding useless detailed studies and tests in connection with installations which simple analyses would show to be relatively innocuous, that is, to be in category I. After an installation in this category is placed in service, a few tests may be made to check the indications of the rough estimates and catalogue the case for future reference.

2. It will be easy to recognize those installations in which the only way to find out definitely whether remedial measures are needed—and if so the best type—is by detailed tests of power system influence and telephone noise before and after the rectifier is put into operation, that is, installations in category II.

3. The occasional situation in which the rectifier load would cause immediate and severe problems, that is, installations in category III, can be recognized and suitable precautions taken. These precautions are, usually, to provide remedial measures in advance, or if this is not feasible, to confine the harmonics to areas where they can do little harm until tests can be made and remedial measures designed and installed.

4. It will facilitate co-operative advance planning, for example, selection of methods of power supply and number of phases, to minimize the probability of important induction problems.

5. Where tests are required, it will facilitate setting up test programs to obtain information of the maximum usefulness with minimum effort.

While this report is devoted primarily to rectifiers of the sort usually encountered in industrial, railway, and radio applications, methods are given later for treating certain of the types encountered less frequently. The following matters pertaining to harmonics associated with rectifiers are not covered, but references to other information are included as indicated:

1. Harmonics on the d-c side of rectifiers (see items 21, 32, 33, 34, 35, and 36 in References).

2. Effect of harmonics on the a-c side on heating of power equipment and on balance of rectifiers (see items 3, 16, and 31 in References).

3. Single-phase rectifiers (see item 36 in References).

More complete treatments of specific aspects of the subject are contained in the literature listed under References, which are referred to in the following text. Appendixes I and II contain, respectively, brief discussions of the fundamentals of induction and of the propagation of harmonic currents and voltages over power circuits, which may be found useful in obtaining a more comprehensive picture of the phenomena involved. Appendix III gives pertinent data on a few of the installations which have been studied.

Nature of Problem

In this section are discussed the basic factors of the inductive co-ordination problem relating to rectifier installations from the point of view of the a-c power system. An understanding of these factors will be helpful in analyzing situations involving inductive co-ordination problems and in finding remedial measures.

Factors Affecting Problem

Considering the a-c side of a rectifier installation, the only significant effect of the installation from the inductive co-ordination standpoint is the production of harmonic currents and voltages in the power system.[1, 2, 3, 8, 9, 17, 30, 31, 32, 34, 53] These harmonic currents and voltages set up magnetic and electric fields of corresponding frequencies about the power lines; they, in turn, may induce voltages of the same frequencies in paralleling telephone circuits which may cause noise in these circuits. Induction caused by power circuit currents is called magnetic; that caused by power circuit voltages is called electric (see Appendix I). Whether noise problems will occur and, if so, their severity and extent, will depend on the magnitudes of the harmonic currents and voltages at various places over the power system, the characteristics of the power system, the amount, type and location of situations of proximity between power and telephone circuits (that is, "exposures"), and the characteristics of the telephone circuits. These matters are discussed more fully in Appendix I.

Generation of Harmonics

The current taken by a rectifier unit from the a-c line has a step-type wave shape resulting from the flat top wave shape of the anode currents. Typical wave shapes of rectifier a-c line currents are shown in Figure 1.

Table I shows the orders (that is, the multiples of the fundamental frequency) of the harmonics on the a-c circuits associated with rectifiers of different numbers of phases. The magnitude of any harmonic which is present with any particular number of phases is the same as for a 6-phase rectifier if all other conditions are the same. For example, the magnitudes of the twenty-third and twenty-fifth harmonics are approximately the same whether the rectifier has 6, 12, or 24 phases.[3, 5, 8, 9, 20, 31]

The magnitudes of the harmonic currents have a definite relation to the magnitude of the total rectifier current and also depend on the order of the harmonic. The higher the order of a harmonic, the smaller is its magnitude.

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Effect of Power Circuit Impedance

In Figure 2 is shown a simple power supply circuit for a rectifier installation. The rectifier unit R is supplied from a power source G over a 3-phase line A. The power supply circuit has an impedance Z, which includes the impedances of the line and the power source. From the point of view of the power circuit, the rectifier unit can be considered as a generator of harmonic currents. The impedance of the power supply circuit will effect some reduction in the harmonic currents, the extent of the reduction depending on the magnitude of the impedance.

If a harmonic current I_n flows from the rectifier over the line to the power source, a voltage drop in the line will result because of the impedance Z_n of the power circuit at the harmonic frequency. This will produce at point S a harmonic voltage $E_n = I_n Z_n$. If a communication circuit should have an exposure to the power line A, experience shows that the controlling inductive effect in such a case is almost always caused by the harmonic current rather than by the harmonic voltage; the latter can, therefore, be neglected without much error for an exposure to the main supply feeder.

However, if the power line should have an extension B for supplying other loads, as shown in Figure 3, the harmonic voltage at S will cause a harmonic current to flow over that line also, although the power to the rectifier is supplied only over line A. This harmonic current in, as well as the harmonic voltage on, line B and any lines that may branch from it could cause induction in communication lines exposed to them.

The higher the value of impedance Z_n the greater will be the harmonic voltage at S and the higher the magnitude of the harmonic current flowing over line B. Actually, the harmonic currents from a rectifier can flow into any part of an a-c system to which it is connected, as determined by the impedances of the various branches of the system at the harmonic frequencies, regardless of the location of the source of power supply or the direction of power flow.[18,19] In practice, however, the circuit over which the rectifier receives its power usually has the lowest impedance and draws the greater part of the harmonic currents.

Table I. Harmonics Arising in Rectifiers

Orders of Harmonics Present With Balanced Operation*						
6-Phase Rectifier	12-Phase Rectifier	18-Phase Rectifier	24-Phase Rectifier	36-Phase Rectifier	48-Phase Rectifier	Harmonic Frequencies on 60-Cycle System, Cycles
5	—	—	—	—	—	300
7	—	—	—	—	—	420
11	11	—	—	—	—	660
13	13	—	—	—	—	780
17	—	17	—	—	—	1,020
19	—	19	—	—	—	1,140
23	23	—	23	—	—	1,380
25	25	—	25	—	—	1,500
29	—	—	—	—	—	1,740
31	—	—	—	—	—	1,860
35	35	35	—	35	—	2,100
37	37	37	—	37**	—	2,220
41	—	—	—	—	—	2,460
43	—	—	—	—	—	2,580
47	47	—	47	—	47	2,820
49	49	—	49**	—	49**	2,940
53	—	53	—	—	—	3,180
55	—	55**	—	—	—	3,300
59	59	—	—	—	—	3,540
61**	61**	—	—	—	—	3,660**

* Other harmonics may be present if the rectifier is unbalanced, as by differences in load or phase control between units, or by harmonics in the power supply.

** The series continues above these values.

Effect of A-C Line Voltage

For a given kilovolt-ampere input to a rectifier installation, the higher the a-c line voltage the lower is the a-c line current and the lower the magnitude of the harmonic currents, because the magnitude of the harmonic current components has a definite relation to the rms value of the rectifier current. When the harmonic currents are lower, usually the harmonic voltages produced by them are also lower. Also, there are fewer long close exposures of telephone circuits to high-voltage transmission circuits than to power distribution circuits. Thus as the voltage of the line supplying a rectifier installation is increased, the probability of interference in communication circuits in the vicinity of the rectifier installation is decreased.

Harmonic currents and voltages can be transmitted through power and distribution transformers in the a-c system and their magnitudes changed in accordance with the transformation ratios of the transformers. However, the factors controlling the propagation of harmonics over power networks are different quantitatively from those controlling the fundamental, as discussed in more detail in Appendix II.

Effects of Phase Control

Phase control is used on a rectifier for the purpose of reducing the d-c output voltage below the value obtained without phase control. It is accomplished by retarding the firing point of the anodes in the alternating-voltage cycle, through grid or firing control. When phase control is used, there is no change in the transformation ratio of the rectifier transformer, and the ratio between the rectifier d-c current and the a-c line current remains practically unchanged.

If the d-c output is maintained constant while the d-c output voltage is reduced by phase control, the kilowatt output of the rectifier unit is reduced while the magnitude of the a-c line current and the kilovolt-ampere input remain unchanged, the power factor at the a-c line terminals of the rectifier transformer being lowered thereby. The result is that the magnitude of the harmonic components in the a-c line current, per kilowatt output of the rectifier, is increased.

Phase control also produces a steeper rise and fall of the anode currents at the beginning and the end of the anode firing period, while the current is commutated between successive anodes (see Figure 4). This is reflected in a steeper change of current between successive steps of the a-c line current wave, which results in harmonic components of higher magnitude.[8,9]

These effects increase as the amount of phase control employed is increased. The use of an appreciable amount of phase control is therefore likely to increase the co-ordination problem between a power system supplying rectifiers and communication circuits. It is important, therefore, that the use of phase control in excess of actual requirements be avoided, particularly in the range of full-load and overload ratings of rectifiers.

I·T Product and Voltage TIF

In the remainder of the body of the report, the terms "I·T product" and "voltage TIF" are used freely. These terms are discussed in Appendix I. At this point it is sufficient to note that I·T prod-

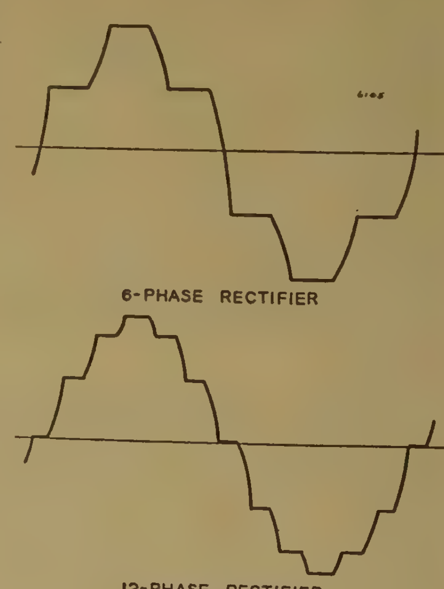


Figure 1. A-c line currents

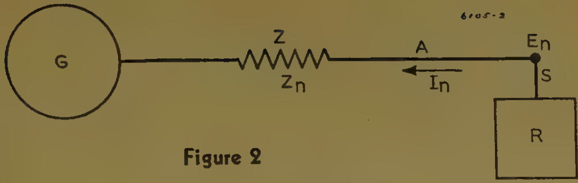


Figure 2

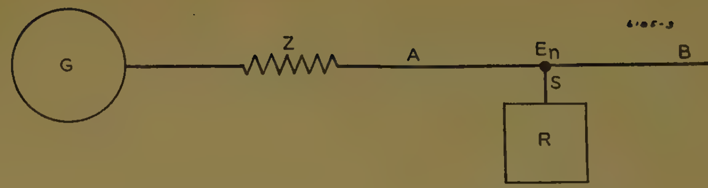


Figure 3

uct is a measure of the over-all noise influence of a current (including fundamental and harmonics); voltage TIF is a similar measure for voltage of known rms value.[27,28]

Ground Return and Balanced Harmonics

Harmonics arising in rectifiers supplied on a 3-phase basis start out as balanced, that is, the phase currents are equal and 120 degrees apart, as are also the phase-to-neutral voltages. If the power system were perfectly balanced, that is, if at harmonic frequencies the series impedances of the three phase wires were equal and also if the admittances to ground of these wires were equal, these currents and voltages would remain balanced. However, with three or more wires on a pole head, the series impedances of the different wires are usually not identical nor are the admittances to ground identical. In many cases these inequalities are accentuated by the presence of single-phase branches and single-phase loads. As a result, on distribution circuits (and to a limited extent on transmission circuits), the vector sum of the harmonic phase currents or of the phase-to-ground voltages is not zero; the vector sum is called "residual." That part of the residual harmonic current which returns in the ground is called "ground return"; it has a larger inductive effect than currents confined to the wires. Similarly, the effect of residual voltages is larger than that of balanced voltages. These matters are discussed in more detail in Appendix I.[10,11,12,23]

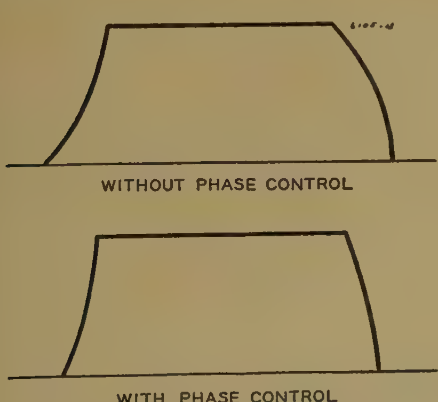


Figure 4. Anode currents

Relations Between Harmonic Voltage and Harmonic Current

Harmonic voltages on a power system may be important from two standpoints, namely:

1. Where open wire telephone circuits are involved, the induction from these voltages may be important.

2. In many situations, the more important effect of harmonic voltages is in causing harmonic currents in parts of systems on which these voltages are impressed. For example, in Figure 3 the effect of a harmonic voltage E_n in causing a harmonic current in part B of the system is likely to be more important than the direct effect of E_n. This is particularly likely to be true if part B is unbalanced, as by single phase branches and loads, so that part of the current becomes ground return.

In connection with 1 above, the harmonic current into B at any frequency f_n is the impressed harmonic voltage of that frequency E_n divided by the impedance looking into B at that frequency. This impedance is not the series impedance of the wires of B alone; it depends not only on these series impedances but also on the impedances in shunt with the circuit, such as, for example, the impedances of loads or capacitors connected to the circuit and of the capacitances of the wires to each other and to earth. All of these impedances vary with frequency—some directly, some inversely, and some in a complex manner. Estimating the I·T on an actual circuit under these conditions would, therefore, be quite involved. For rough estimates only, it has been found useful to assume a direct relation between the voltage TIF on the circuit and ground return I·T.

Procedure in Classifying Proposed Installations

On pages 420 and 421 are shown the steps involved in the classification of a proposed rectifier installation together with essential data and equations. In the remainder of this section, these steps and data are discussed.

Step 1 (Refer to Table III, Page 420)

It is convenient to start with a schematic diagram of the power supply system. A typical diagram, shown in Figure 5A, includes the rectifier and its transformer, the supply feeder and its transformer, and associated power sources which may involve generators, a supply network, or both. In addition, the location of present or proposed telephone circuits, in relation to power circuit feeders, should be determined as illustrated by exposures to power feeders A, B, C, and N. These exposures illustrate the groupings that are convenient in classifying installations from the co-ordination standpoint. While in a detailed study it is necessary to have rather complete information on exposure conditions, it usually is sufficient for classification purposes to know their general type and extent; for example, the extent to which the power circuits are or are likely to be involved in joint use of poles with telephone cables or in joint use or roadway exposures to telephone open wire.

It is usually convenient, next, to draw an impedance diagram similar to Figure 5B. In this diagram the reactances of the various parts of the circuit through which power flows to the rectifier are indicated as percentages on the total rectifier transformer kilovolt-ampere base. If the rectifier transformer kilovolt-amperes is known, it is used directly; if it is not known, a sufficiently good approximation for the present purpose can be obtained from equation 1. The per cent commutating rectance ($\%X_{cn}$) of the rectifier transformer can be obtained from the manufacturer or can be approximated from Table II.*

* The per cent impedance given on rectifier transformer nameplates should not be used because it applies only to the case where voltage is applied to the primary, and all of the secondary terminals are short-circuited.

Table II. Approximate Per Cent Commutating Reactance of Rectifier Transformers

	Approximate Reactance, Per Cent	
Voltage Class on A-C Side, Kv	Rectifiers for Railway and General Industrial Service	Rectifiers for Electro-chemical Processes
Up to 15	5	9
34.5	6	9
69	8	9
115	10	10
161	12	12
230	14	14

Table III. Summary of Classification Procedure

Step 1. Schematic and Impedance Diagrams of Power Supply System (as Figure 5)

1. Indicate exposures
2. Classify circuits as to type (as *A*, *B*, *C*, or *N*)
3. Determine per cent reactance of circuit elements at 60 cycles on total rectifier transformer kilovolt-ampere base
   - *A*. Commutating reactance of rectifier transformer from Table II or from manufacturer
   - *B*. 60-cycle power source reactance (use name plate data or Tables VIII and IX for reactances, and equations 4 and 5 for conversion to rectifier transformer kilovolt-ampere base)
   - *C*. Rectifier feeder (see equations 2 and 3)
4. Obtain *total* per cent commutating reactance (total $\%X_c$) on total rectifier transformer kilovolt-ampere base by combining items 3A, 3B, and 3C above; or
5. If data on reactances are not available, assume total per cent commutating reactance of eight per cent for industrial and railway installations supplied from distribution circuits and the values in Table II plus two per cent for all other installations

Step 2. Direct-Feeder Exposure (Exposure A, Figure 5)

1. If phase control is not used at rectifier rated load, estimate I·T product from Figure 6
   - *A*. Interpolate between curves, using total $\%X_c$ obtained in item 4 or 5 of step 1
   - *B*. Divide ordinate by phase-to-phase voltage in kilovolts and multiply by total rectifier transformer kilovolt-amperes
2. If phase control is used at rated load, obtain I·T product as discussed under "Substantial Amounts of Phase Control" rather than from Figure 6
3. Classify exposure as in Table IV*

Step 3. Indirect Distribution Exposures (Exposure B and C of Figure 5)

1. If phase control is not used at rectifier rated load, estimate voltage TIF on bus supplying feeder *B* or *C*, from Figure 7
   - *A*. Interpolate between curves, using total commutating reactance ($\%X_c$) obtained in item 4 or 5 of step 1
   - *B*. Multiply ordinate by per cent reactance from (and including) 60-cycle power source to point of power supply to feeder *B* or *C*
2. If phase control *is* used at rated load, obtain voltage TIF as discussed under "Substantial Amounts of Phase Control" rather than from Figure 7
3. Classify exposure as in Table V*

Step 4. Exposures to Supply System Network (Exposure N, Figure 5)

1. Starting with the I·T product obtained in step 2
   - *A*. Obtain division of this I·T product between network (*N*) and generator (*G*) on basis of their inverse impedances
   - *B*. Multiply *network* I·T from (*A*) above, by line voltage of direct feeder (*A*) and divide by line voltage of the *network*
2. Classify exposures as in Table VI*

Estimates of per cent reactance of other circuit elements on total rectifier transformer kilovolt-ampere base can be obtained from equations 2, 3, 4, and 5. The following suggestions may be of help in estimating reactances to use in these equations:

1. For generators and other synchronous

* In Tables IV, V, and VI, the classification of installations into categories has the following general significance:

Category I. Very *unlikely* to cause important induction problems.

Category II. Indeterminate.

Category III. Very *likely* to cause important induction problems.

Table IV

Type of Feeder A	Type of Telephone Plant	Approximate I·T Product on Feeder A for Category I*	Category II*	Category III*
Cable	All	All	—	—
Overhead circuit, but so located that there are and will be no exposures more than a few spans long	All	All	—	—
Overhead distribution circuits with normal existing or possible future exposures	Cable†	Less than about 10,000	About 10,000 to about 100,000	More than about 100,000
	Long open wire	Less than about 5,000	About 5,000 to about 50,000	More than about 50,000
Rural,‡ with present or possible future exposures	Long open wire	—	All	—

† With or without short open-wire extensions. ‡ Feeder of low capacity with single-phase loads.

Table V

Type of Distribution Circuit Involved in B and C	Type of Telephone Plant	Approximate Voltage TIF on B or C for Category I*	Category II*	Category III*
Cable	All	All	—	—
Overhead, so located that no exposures more than a few spans can exist	All	All	—	—
Overhead, urban, with no capacitors more than a few blocks away from the rectifier	Cable†	Less than about 50	About 50 to about 150	More than about 150
Overhead, urban with a likelihood of capacitors some distance away	Cable or open wire	Less than about 40	About 40 to about 100	More than about 100
Overhead circuit supplying scattered industrial and residential loads in relatively open country and without extensive single-phase branches	Long open wire	Less than about 30	About 30 to about 75	More than about 75
Rural‡	Long open wire	Less than about 20	About 20 to about 50	More than about 50

† With or without short open-wire extensions. ‡ Feeder of low capacity with single-phase loads.

Table VI

Characteristics of Supply System	I·T Product Into the System (N) Category I*	Category II*	Category III*
Extensive cable networks of type found in large cities	All	—	—
Overhead subtransmission systems	Less than about 20,000	About 20,000 to about 100,000	Over about 100,000
Extensive high-voltage transmission systems:			
A. Rectifier with 6–24 phases	Less than about 5,000	5,000 to 25,000	Over 25,000
B. Rectifier with 30–48 phases	Less than about 10,000	10,000 to 50,000	Over 50,000
C. Rectifier with 72 or more phases	All	—	—

Table VII. Equations Used in Estimating Reactances Associated With Rectifier Installations

Nature of Reactance	Equation	Number
To estimate: A-c line kva of rectifier transformer $(\text{kva})_R$ from rectifier kw and phase control	$\text{kva}_R = 1.15\,\frac{(100 + \%\text{ phase control})}{100} \times \text{rectifier kw}$ Here "% phase control" is amount of phase control at rated voltage and load	(1)
Per cent reactance ($\%X_L$) of line of length L (in miles), on rectifier transformer kva base	$\%X_L = (0.00375)(L)(\text{kva}_R)$ for 4-kv circuits	(2)
	$\%X_L = \frac{(0.075)(L)(\text{kva}_R)}{E_L^2}$ for 13-kv and higher-voltage circuits, where E_L is phase-to-phase voltage, in kv	(3)
To find: Per cent reactance ($\%X_{Rz}$), on rectifier transformer kva base (kva_R), of equipment having a known per cent reactance ($\%X_z$) with respect to some other kva base (kva_z)	$\%X_{Rz} = \%X_z\,\frac{\text{kva}_R}{\text{kva}_z}$	(4)
Per cent reactance ($\%X_{Rz}$), on rectifier transformer kva base (kva_R), of a reactance given in ohms (X_z) in a line of voltage E_L	$\%X_{Rz} = (X_z)\,\frac{(\text{kva}_R)}{10E_L^2}$ E_L is phase-to-phase voltage, in kv	(5)

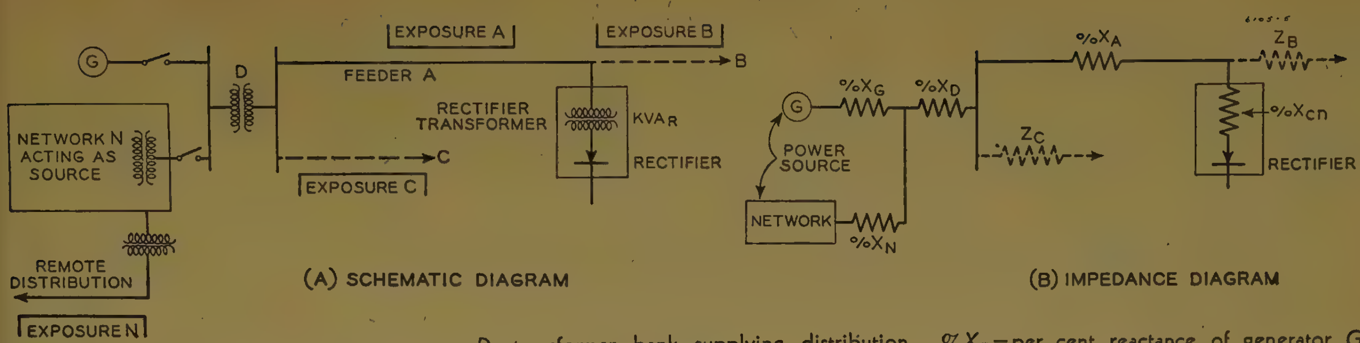


Figure 5. Typical power-supply and telephone exposure diagram for classifying rectifier co-ordination problems

A = feeder to rectifier

B = distribution system extending beyond rectifier station

C = distribution system fed from supply substation

D = transformer bank supplying distribution system

G = generator

$\%X_{cn}$ = per cent commutating reactance of rectifier transformer

$\%X_A$ = per cent reactance of feeder A, on rectifier transformer kilovolt-ampere base

$\%X_D$ = per cent reactance of transformer D, on rectifier transformer kilovolt-ampere base

$\%X_G$ = per cent reactance of generator G, on rectifier transformer kilovolt-ampere base

$\%X_N$ = per cent reactance of network N, on rectifier transformer kilovolt-ampere base

Z_B = impedance presented by distribution circuit B, to voltages of nth harmonic frequency

Z_C = impedance presented by distribution circuit C, to voltages of nth harmonic frequency

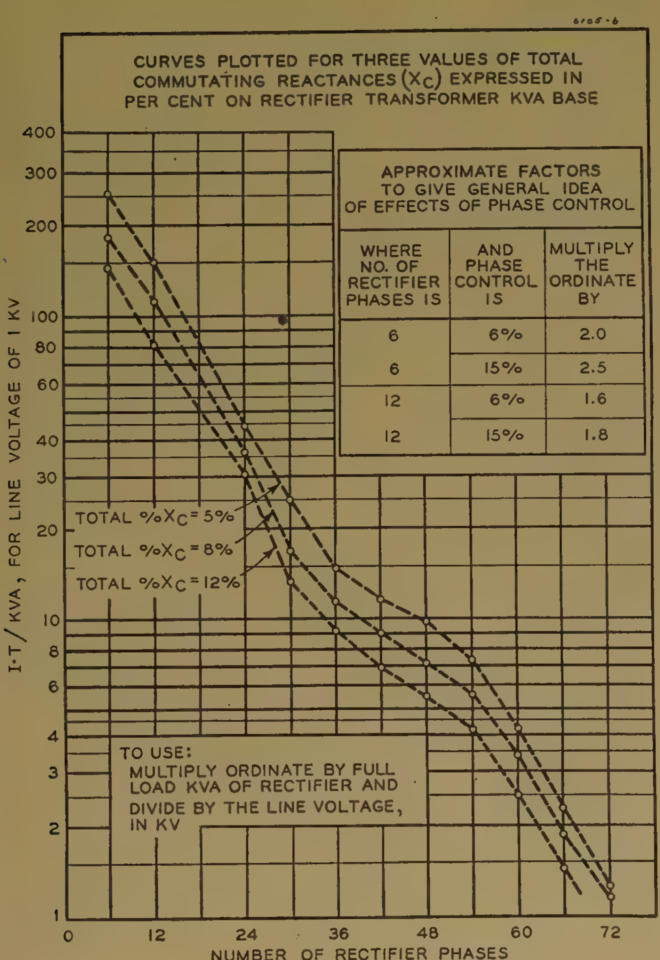


Figure 6. I·T per kilovolt-ampere of rectifier transformer capacity for 1-kv line voltage versus number of rectifier phases (60-cycle systems)

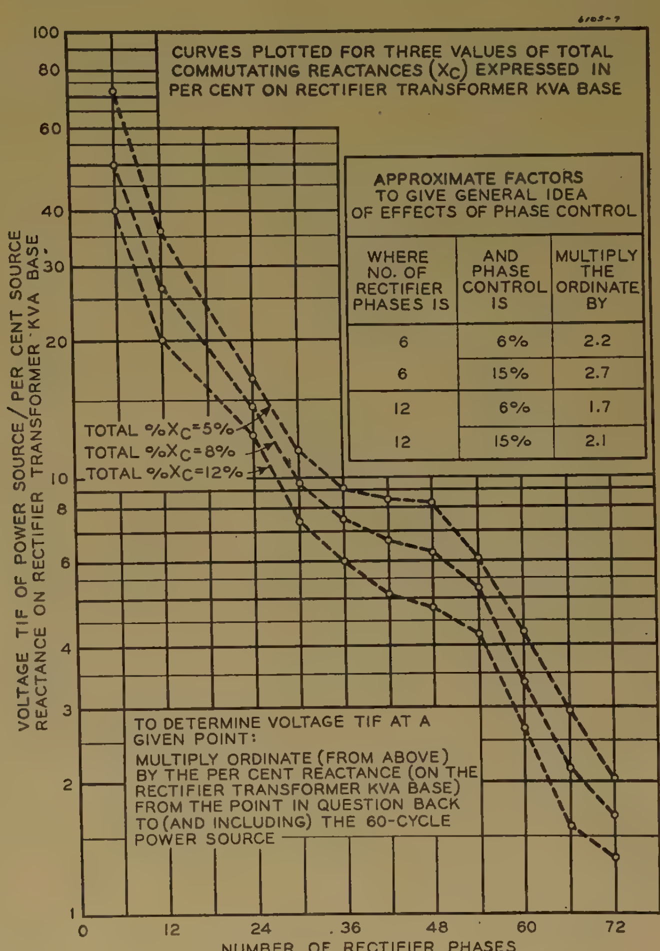


Figure 7. Source voltage TIF per per cent source reactance on rectifier transformer kilovolt-ampere base versus number of rectifier phases (60-cycle systems)

machinery, the average of the subtransient and negative sequence reactances may be used. If these reactances are not known, the approximate percentages on the machine kilovolt-ampere base, shown in Table VIII,[14,37] may be used.

2. If the reactance of a power transformer bank is not known, the approximate figures in Table IX may be used.

Table VIII. Approximate Reactances of Synchronous Machines

Type of Machine	Reactance on Own Base for Use in Equation 4, Per Cent
2-pole turbine generator	9
4-pole turbine generator	15
Salient pole generator or motor	
With dampers	24
Without dampers	45
Synchronous condenser	25

3. Where the "network acting as power source" includes a transformer bank between the lines and the bus, about the best approximation is to consider the impedance, looking into it from the bus as being equal to the reactance of the transformer bank alone. This approximation is good if the network is

Table IX. Approximate Reactances of Power Transformers

Voltage Class on High-Voltage Side of Transformer, Kilovolts	Approximate Reactance on Own Base for Use in Equation 4, Per Cent
Up to 15	5
34.5	7
69	8
115	10
161	12
230	14

a high voltage system; it gets successively poorer as the voltage of the network approaches that of the bus.

Having set down the reactances as percentages on the total rectifier transformer kilovolt-ampere base, they are combined to obtain the approximate total commutating reactance in per cent on rectifier transformer kilovolt-ampere base, that is, total $\%X_c$.* In cases where data are not available on reactances, total $\%X_c$ of eight per cent can be used for industrial and railway installations supplied from distribution systems and the values shown in Table II plus two per cent for all other installations.

Step 2. Direct Exposures to Rectifier Feeders (Exposure *A* in Figure 5)

Step 2 covers a direct distribution type feeder supplying power to a rectifier. As indicated under "Nature of Problem," experience indicates that in practically all such cases, magnetic induction is controlling, and that electric induction can be safely neglected. While the ground return I·T is the most significant measure of magnetic influence, it is difficult to estimate; however, for given power circuit conditions, the ground return I·T is about proportional to phase I·T, and the latter provides a sufficiently reliable indication for classification purposes.

It has been found for installations of ordinary rectifiers of a given number of phases operated at full load without phase control, that the quantity "I·T product per kilovolt-ampere of rectifier transformer capacity" usually lies within a fairly narrow range. In Figure 6 are plotted average values of "I·T product per kilovolt-ampere of rectifier transformer capacity" for a 60-cycle system having a phase-to-phase voltage of one kv. Three curves are shown for values of total commutating reactance (total $\%X_c$) of 5%, 8%, and 12% on the rectifier transformer kilovolt-ampere base. The total commutating reactance at rated load may be determined as explained previously and then used to interpolate between the three curves of Figure 6. The I·T per kilovolt-ampere thus obtained, when multiplied by the rectifier transformer kilovolt-amperes and divided by the a-c phase-to-phase voltage in kilovolts, will give the I·T product for feeder *A*. The data from Figure 6 also apply if phase control is used only at fractional load or if small amounts of phase control (not over a few per cent) at full load are used to balance loads among units forming a multiphase installation.** If larger amounts of phase control are used in normal operation, the methods outlined in "Installations Using Substantial Amounts of Phase Control" should be used.

With known I·T product, installations can be classified from the standpoint of exposures similar to exposure *A* as in Table III.

Step 3. Indirect Distribution Exposures (Exposures *B* and *C* of Figure 5)

From the standpoint of exposures *B* and *C* in Figure 5, the factor of primary interest usually is the ground return I·T product. This factor, however, is very difficult to estimate, and for that reason, the voltage TIF impressed on the feeders and their extent and type are usually employed. This approximation is rather poor quantitatively but is good enough for classification purposes. Where open wire telephone circuits are involved, the voltage TIF itself is of importance.

The voltage TIF on feeder *C* at the substation is produced by the voltage drops caused by the harmonic currents flowing through the reactances of transformer bank *D* and the source as explained in "Nature of Problem." The voltage TIF may be estimated with the aid of Figure 7. The value of the total commutating reactance obtained in step 1 is used to interpolate between the curves of Figure 7. The ordinate thus obtained for the particular number of rectifier phases is the voltage TIF contribution per "per cent power source reactance" on the rectifier transformer kilovolt-ampere base. This value when multiplied by the appropriate per cent power source reactance gives the voltage TIF on the feeder. "Power source reactance," as here used, is the reactance on the power source side of the point considered. In case of feeder *C*, the power source reactance includes the reactances of transformer *D* and of the power supply to it.

In thus using Figure 5, the estimated voltage TIF impressed on feeder *B* is the voltage TIF per per cent source reactance as estimated above times the power source reactance to the point where feeder *B* is supplied. The per cent source reactance to the point where feeder *B* is supplied is equal to the source reactance used for feeder *C* plus the per cent reactance of feeder *A*. The reactance of feeder *A* can be determined conveniently from equation 2 or equation 3, depending on the circuit voltage.

Having obtained estimates of voltage TIF at the points supplying power to feeders *B* and *C*, installations can be classified from the standpoint of exposures *B* and *C* as in Table V.

If no exposures exist or will be created in the future in a particular location, there is no occasion for estimating voltage TIF for that feeder location or supply bus.

Step 4. Exposures to Supply System Network (Exposure *N* in Figure 5)

The power supply system may range from a single generator *G* to many generators and may include an extensive transmission or subtransmission network. Where such a network is involved, the primary consideration from an induction standpoint is whether the harmonic current fed to it will have a significant effect on the influence of distribution systems (particularly rural) connected to it and over how large an area this will occur. The rectifier harmonics do, of course, tend to increase the influence of the high voltage circuits themselves, but as a rule problems from this cause are less widespread or important than problems associated with the effects on distribution systems supplied by these circuits. Experience has shown that the most useful factors in estimating the effect of rectifier harmonics from this standpoint are:

1. The I·T product put into the network by the rectifier.

2. The general characteristics of the network.

The I·T product on the direct feeder *A* has already been obtained in step 2. This I·T product will divide between the generator and network circuits in the inverse ratio of their impedances, from which the I·T product put into the network can be estimated. Where the network is connected to the bus through a transformer bank, the impedance looking into the

* For rectifier installations of 12 or more phases, strict theoretical accuracy would require application of correction factors to the a-c system reactances before combining them with the commutating reactance of the rectifier transformer to obtain total $\%X_c$. For the purposes of this report, this refinement was not considered essential and was omitted for the sake of simplicity.

** It has been found by experience that the minimum I·T is secured when the loads among units in a multiphase installation are well balanced, even if this requires the use of small amounts of phase control in some of the units.

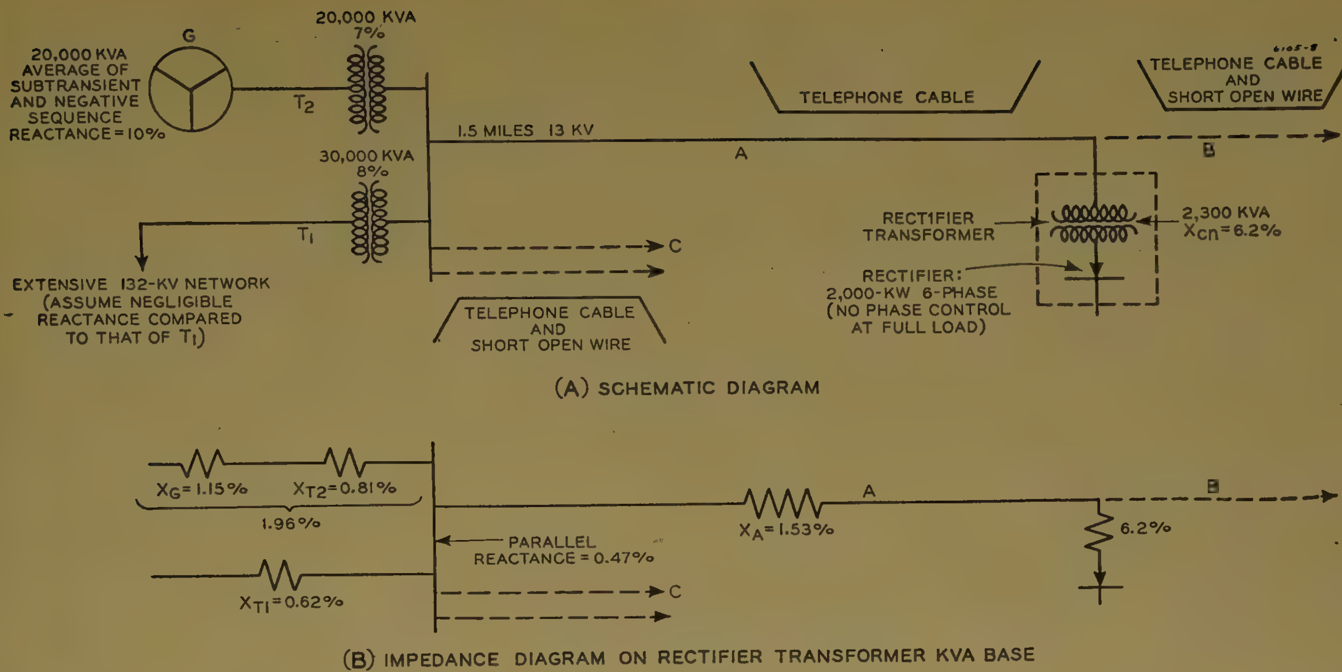


Figure 8. Illustrative example of installation to be classified

network can be assumed to be that of the transformer bank; where there is no transformer bank, the safest assumption is that all of the I·T goes into the network. In estimating magnitude of I·T products at various points, it is, of course, necessary to take into account the ratios of any transformers through which the currents pass.

While a wide range of supply network and exposure conditions may be encountered, the most useful and convenient classification from the co-ordination standpoint is based on the information outlined in Table VI.

Discussion

Rectifier co-ordination problems involve complex phenomena. As a consequence, a few installations which appear to be in category I cause important problems; conversely, a few installations which appear to be in category III turn out to be innocuous. Categories I and III have been set up so as to make them as definite as practicable, while category II covers a wide range of intervening conditions. The chance of a particular case's actually falling outside of its indicated range depends to a considerable extent upon its nearness to the edge of the range.

Certain additional factors may be considered qualitatively in estimating the likelihood of important induction problems for installations, particularly those in category II. Some favorable and unfavorable factors in exposures to distribution circuits (exposures A, B, and C of Figure 5) are:

1. Heavy industrial type feeders supplying mainly 3-phase loads provide a favorable factor. The lighter the load, the more scattered it is; and the larger the proportion of single-phase line and load, the greater will be the likelihood of a co-ordination problem, assuming normal urban exposure conditions.

2. Large banks of capacitors at or near rectifier installations of 12 or more phases are usually favorable; for 6-phase installations they may be favorable or unfavorable. The reason for the differentiation is that resonance between capacitors and urban distribution circuits is most likely to be important for frequencies of from 200 to 500 cycles, which are normally suppressed with rectifiers of 12 or more phases.

3. Capacitor banks electrically remote from rectifier installations may be unfavorable, particularly for 6-phase rectifiers.

Certain additional favorable and unfavorable factors in exposures involving the supply network (exposure N of Figure 5) are:

1. For urban power cable networks, the greater the lengths of cable which the harmonics must traverse before reaching overhead circuits, the less is the likelihood of a problem.

2. The larger the number of transformers through which the harmonics must pass before they reach overhead distribution circuits, particularly rural circuits, the smaller is the likelihood of a problem.

3. Supply to a rectifier direct from a generator bus is favorable; a generating station connected to a high voltage transmission system through transformers may be favorable or unfavorable; usually the effect is not large.

In a number of the cases investigated, it was found that the rectifiers were operating with unnecessarily large amounts of phase control at full load. As a result, the I·T product and voltage TIF were substantially higher than would normally be expected. In most of these cases, selection of different transformer taps and, occasionally, readjustment of the phase-control mechanism reduced the harmonics substantially.

In some installations, one or more of circuits A, B, or C may be missing and need not be considered in the classification procedure. For example, some installations are supplied from high voltage networks through transformer banks which supply power only to the rectifiers and the plants in which they are located; in such cases only the I·T into the high voltage network is of interest.

Illustrative Example

In order to illustrate the application of these methods, consider Figure 8.

Step 1. Figure 8A shows a schematic diagram of an assumed power system associated with a rectifier, and Figure 8B shows an impedance diagram on the rectifier transformer kilovolt-ampere base.

Assume that the exposures to A, B, and C are normal urban, miscellaneous joint use and roadway, and that power circuits B and C extend into residential areas.

The reactances in Figure 8B were determined as follows:

$$\%X_A = \frac{0.075 \times 1.5 \times 2{,}300}{13^2}$$

$$= 1.53\% \text{ (using equation 3)}$$

$$\%X_{T_1}=8\times\frac{2{,}300}{30{,}000}$$
$$=0.62\% \text{ (using equation 4)}$$

$$\%X_{T_2}=7\times\frac{2{,}300}{20{,}000}$$
$$=0.81\% \text{ (using equation 4)}$$

$$\%X_G=10\times\frac{2{,}300}{20{,}000}$$
$$=1.15\% \text{ (using equation 4)}$$

$$\%X_G+\%X_{T_2}=1.15+0.81=1.96\%$$

Reactance of T_1 and generator branch in parallel is

$$\frac{1}{\frac{1}{1.96}+\frac{1}{0.62}}=\frac{1}{0.51+1.61}=\frac{1}{2.12}=0.47\%$$

The total commutating reactance on rectifier transformer kilovolt-ampere base:

$$\text{Total } \%X_c=6.2+1.53+0.47=8.20\%$$

Step 2. The total I·T in feeder A is determined as follows:

With a total X_c of 8.2 per cent, the I·T per kilovolt-ampere on a 1-kv circuit is, from Figure 6, about 180. The total I·T in A is then:

$$\frac{180\times 2{,}300}{13}=32{,}000$$

From Table IV, the installation would be classified in category II from the standpoint of the direct feeder A.

Step 3. The per cent power source reactance used to estimate the voltage TIF on feeders C is the reactance of the generator branch and T_1 in parallel, which is 0.47%. From Figure 7, the voltage TIF per per cent source reactance is (for a total X_c of 8.2%) about 50. The voltage TIF impressed on C will thus be about $50\times 0.47=24$.

From Table V, the installation would be classified in category I, from the standpoint of feeders C.

The per cent power source reactance used to estimate the voltage TIF impressed on B is $0.47+X_A=0.47+1.53=2.00\%$. Again, the voltage TIF per per cent is 50; the voltage TIF on B will then be about $50\times 2.0=100$.

From the standpoint of circuit B, the installation will be (from Table V) in category II. If capacitors are to be used on the power system, it will be near the upper edge of category II. If circuit B should be used to supply a rural system, the installation would probably be in category III from the standpoint of the rural system.

Step 4. The total I·T on a 13-kv base going into the generator and high voltage network branches in parallel is (from step 2) 32,000. That going into the generators is of no interest. That going into the transmission network is

$$32{,}000\times\frac{1.96}{1.96+0.62}=24{,}000$$

On the 132-kv high voltage network, this I·T is reduced by the ratio of transformation of transformer T_1, or it is

$$24{,}000\times\frac{13}{132}=2{,}400$$

From Table VI, it will be seen that from the standpoint of the high voltage network, the installation is in category I.

Discussion of Example

From the analysis, it can be seen that the area in which problems are most likely to occur when the rectifier is placed in operation is that supplied by feeder B. It is possible that there will be some problems associated with feeder A, but unless the exposure extends practically the whole length of the feeder, they probably will not be unduly severe. The chances of problems associated with feeder C or with remote distribution systems fed from the 132-kv network are small.

It would seem wise in this case to examine the situation with regard to feeder B in more detail before placing the rectifiers in operation; for example, ascertain whether capacitors are used at widely separated points, the general degree of circuit balance, and the amount and type of exposure (particularly between the rectifier and the capacitors). Arrangements might be made for applying telephone cable sheath shielding in advance of the start of rectifier operation and for some method of emergency operation to take care on a temporary basis of any difficulties that might develop until permanent measures could be applied if tests indicated their necessity.

It probably also would be advisable, if practicable, to make tests on feeders A and B and on some telephone circuits exposed to them before the rectifier is brought up to full load, possibly during the preliminary tests of the rectifier.

Little, if any, attention need be given to the high voltage network or to feeders C except possibly to make a few tests after the rectifier is in regular operation to be sure that no unforeseen conditions exist.

It is of interest to note that in this case a ten per cent voltage reduction at full load by phase control would result in more than doubling the I·T's and voltage TIF's with a consequent substantial increase in likelihood of problems; for example, it would result in putting the installation well into category III from the standpoint of B, in category II from the standpoint of C, and close to category II from the standpoint of the high voltage network. (See example under "Installations Using Substantial Amounts of Phase Control.")

On the other hand, if the installation were 12 phase without phase control at full load, the I·T's and voltage TIF's would be reduced 35 per cent to 45 per cent, with a consequent substantial reduction in the likelihood of problems.

Small Rectifiers of Three or Four Phases

For small rectifier installations, the capacity may be less than that of 6 or 12 tubes of conventional size. For these cases it may be desirable to use four or eight rectifier tubes with a transformer connected in Scott/4-phase cross, or three tubes with a transformer connected in delta/3-phase zigzag. Estimates of harmonic effects of these connections may be based on the methods of analysis and the data given for 6-phase rectifiers, and increasing the I·T product and voltage TIF factors by the multipliers given in Table X.

Table X

Phase	Relative I·T and Voltage TIF Factors*
4	1.20
3	1.33

* These multiplying factors are to be applied to estimates based on 6-phase rectifiers.

Application to System of Other Than 60 Cycles

In order to apply the methods outlined here in cases where the power system frequency is other than 60 cycles, it will be necessary to obtain voltage TIF and I·T products from other sources.[3,8,9,17] With these factors known, installations can be classified as outlined in the foregoing.

Co-ordinative Measures

The solution of problems involving rectifiers where the application of co-ordinative measures is indicated is no different from other types of inductive co-ordination problems, in that close co-operation of all interested parties is necessary. Guiding principles and general practices are given in the report on "Principles and Practices for Inductive Co-ordi-

nation," by the Edison Electric Institute and the Bell Telephone System,[30] which recognizes that characteristics of both the power and the telephone systems must be taken into account. Of the various kinds of co-ordinative measures applicable to the power or telephone system, the following are the most important in connection with rectifier problems:

1. Advance planning of the method of supplying the rectifier from the standpoint of minimizing wave shape distortion.[32]

2. Phase multiplication of rectifier installation.[3,5,6,30]

3. Frequency selective devices.[21,32]

4. Reduction of power system unbalance to ground.[10,11,12,23]

5. Co-ordinated transpositions.[24] (Transpositions in power distribution circuits are of negligible value but may in some cases be useful in transmission circuits.)

6. Reduction of telephone system unbalance to ground.[10,11,12,23,26]

7. Shielding of telephone cable circuits.[25]

Some of these already have been referred to as factors in evaluating the likelihood of a given installation's causing trouble.

It is impracticable within the scope of this report to present a sufficiently detailed discussion of these measures to provide the information required for reaching a solution of a particular situation. However, the following general discussion of these measures, with some indication of their field of use, may be helpful in a more detailed examination of the material listed in the references.

Induction problems associated with rectifiers are most likely to involve exposures between power distribution circuits and telephone exchange circuits. The difficulties of applying and maintaining co-ordinative measures on a specific exposure basis, where power and telephone distribution circuits are involved, has led in recent years to relying more and more on the maintenance of the lowest practicable susceptiveness* of telephone circuits in general and of the smallest practicable wave shape distortion on power circuits in general. This is particularly true in rural areas where changes and extensions are frequent and where the power circuit influence is very greatly affected by wave shape.

Improved balance-to-ground of the newer types of telephone subscriber and central office equipment, which reached the stage of large-scale production just prior to the war, was most helpful in this connection. Another helpful factor in the prewar years was the increase in use of carrier on telephone toll circuits, which was accompanied by much more effective telephone transposition arrangements. Of great importance from the standpoint of power system influence was the development and application of the principle of phase multiplication, particularly in view of the tremendous expansion during the war in large scale rectifier installations.

Figures 6 and 7 have illustrated the effectiveness of phase multiplication in reducing the I·T and voltage TIF. This results from a reduction in the number of harmonic orders present as indicated in Table I. Another advantage of a relatively large number of phases is that power system resonance conditions tending to magnify the effects of harmonics are much less likely to occur at frequencies above the twenty-fifth harmonic (60-cycle base). Furthermore, if induction problems do arise, remedial measures are much less expensive than for a smaller number of phases because

1. The number of harmonics to be cared for is smaller.

2. The volt-ampere capacity of devices to reduce harmonics goes down rapidly as the lowest frequency which they must handle goes up.

The number of phases in an installation is, of course, limited to the number of anodes. To secure the greatest effectiveness of phase multiplication, the d-c loads on the individual rectifiers must be well balanced. Where they are not well balanced, shifting the phase of successive rectifier units to obtain phase multiplication will be much less effective. When one of the units in a multiphase installation is out of service, as for maintenance, the effectiveness of phase multiplication is reduced, but in only a small number of cases has this presented a problem of importance.

While phase multiplication is a powerful tool in the larger capacity installations, there are still many smaller installations of limited number of anodes where it cannot be employed. The only presently available method of controlling harmonics, other than phase multiplication, is some form of frequency selective device. Most of these devices have consisted of several resonant shunts or of a series reactor and shunt capacitor, per phase. The use of series reactors increases the voltage regulation of the rectifier unit which must be taken into account in many cases. Resonant shunts involve precision tuning and, where several shunts are used per phase, may involve special mounting arrangements.

With the more extensive use of power factor correction capacitors on power systems, a somewhat different approach to the frequency selective device design problem may, in some cases, be advantageous. This approach is first to consider what capacitor installations have been made or can be justified for power reasons and next to consider whether they will, or can be made to, produce harmonic control. For example, if a fairly large block of capacitors can be justified for power reasons at or near the rectifier location, they may be adequate alone, as has been the case in several large metallurgical rectifier installations, or coils to tune the entire bank broadly or to tune parts of the bank to particular frequencies might be added. The same thing might be done with capacitors at a supply substation bus if harmonic currents in the direct feeder are not important. In any such installations—whether equipped with tuning coils or not—the chances of adverse resonance at the lower harmonic frequencies need to be considered. In general, the reduction in I·T and voltage TIF would not be expected to be as large with the broadly tuned arrangement as with several sharply tuned resonant shunts, but in view of its very low cost (assuming the capacitors would be available, anyway), it may be advantageous in some cases, particularly where large reductions are not needed.

The extent to which other measures can be substituted in place of control of the rectifier harmonics in a specific case depends on many circumstances which cannot be described in detail here. However, a few of the more general considerations can be given, namely:

1. Situations where the effect of the rectifier is confined to the feeder between the source of supply and the rectifier offer the best field for the use of measures other than harmonic control, or a combination of these measures with a moderate degree of harmonic control.

2. In cases where the effect of the rectifier spreads beyond the feeder, but only to a limited extent, the same situation exists, although the advantage of reducing the harmonics is greater because of the greater uncertainties as to conditions in the future.

3. In cases where the effects spread over wide areas and involve numerous power distribution systems, it is almost hopeless to attempt to correct induction conditions without limiting the harmonics at the rectifier.

Installations Using Substantial Amounts of Phase Control

For installations employing substantial amounts of phase control at rated load,

* See Appendix I for explanation.

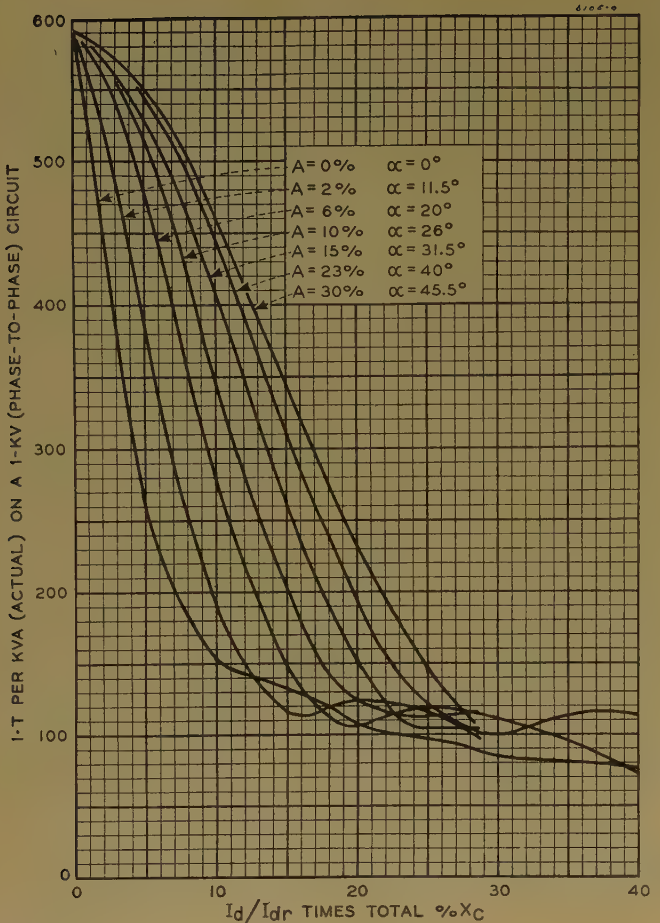


Figure 9. Six-phase rectifiers

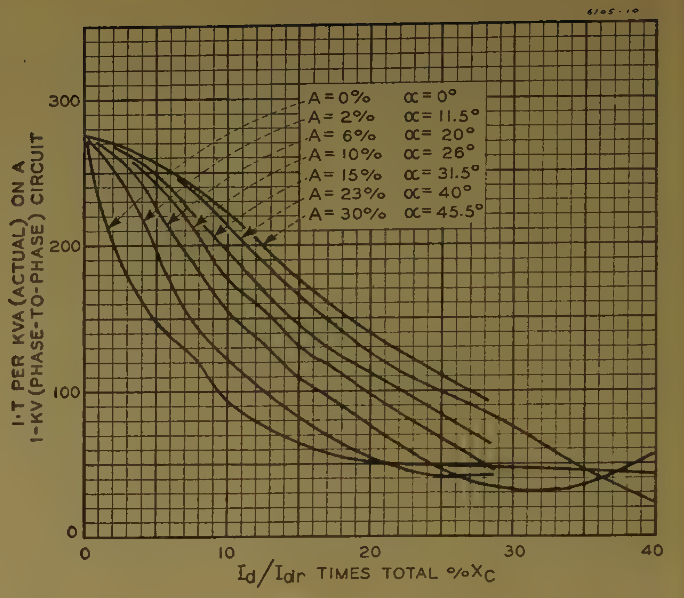


Figure 10. Twelve-phase rectifiers

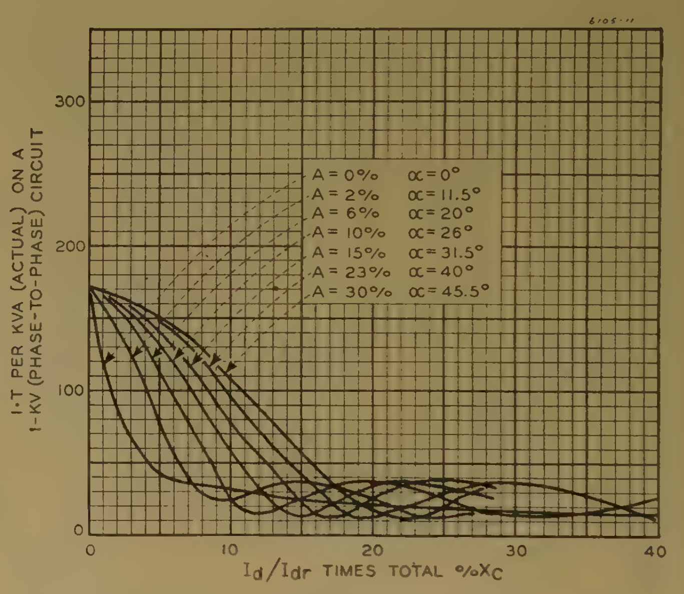


Figure 11. Twenty-four-phase rectifiers

Figures 9, 10, 11. I·T per kilovolt-ampere on 1-kv circuit for 6-, 12-, and 24-phase rectifiers with different amounts of phase control

For 60-cycle systems
I_d = actual d-c load current
I_{dr} = rated d-c load current
Total $\%X_c$ = per cent commutating reactance of rectifier transformer bank and supply system on rectifier transformer bank kilovolt-ampere base
α = phase control angle of retard
A = amount of phase control or per cent voltage reduction below output voltage with zero phase control ($A = 1 - \cos\alpha$)

To find I·T

1. Determine the value of I_d/I_{dr} times total $\%X_c$ for the particular load under consideration. Where the I·T for the rated capacity only is desired, total $\%X_c$ is used directly

2. From the curve for the appropriate number of phases and amount of phase control, determine the I·T per kilovolt-ampere

3. Multiply the I·T per kilovolt-ampere read on the curves by the *actual* kilovolt-ampere and divide by the phase-to-phase line voltage in kilovolts. This is the estimated phase I·T. Actual kilovolt-amperes can be obtained by multiplying rated kilovolt-amperes by the ratio I_d/I_{dr}

Figures 6 and 7 cannot be used to obtain I·T products and voltage TIF's. However, I·T products can be estimated from Figures 9, 10, and 11 which show I·T per rectifier kilovolt-ampere on a 1-kv line with various amounts of phase control, for 6-, 12-, and 24-phase installations, respectively, operating in a balanced manner. Voltage TIF can be estimated by using Figures 12, 13, and 14, which show voltage TIF per per cent power source reactance on rectifier transformer kilovolt-ampere base, with various amounts of phase control, for 6-, 12-, and 24-phase installations, respectively, operating in a balanced manner.

To illustrate the use of these curves, consider the example discussed under "Procedure in Classifying Proposed Installations" (see Figure 8), but assume ten per cent phase control at rated output voltage and current. The I·T product on the direct feeder A and the voltage TIF on B will be estimated:

1. The rectifier transformer kilovolt-amperes required for this rating will be (see equation 1)

$$1.15 \times \frac{(100+10)2{,}000}{100} = 2{,}530 \text{ kva}$$

2. The parallel reactance of the generator and T_1 on the rectifier transformer kilovolt-ampere base will be 0.47 (2,530/2,300) = 0.52 per cent (equation 4).

3. The line reactance (X_A) on rectifier transformer kilovolt-ampere base will be 1.53(2,530/2,300) = 1.7 per cent.

4. The total $\%X_c = 0.52 + 1.7 + 6.2 = 8.4$ per cent.

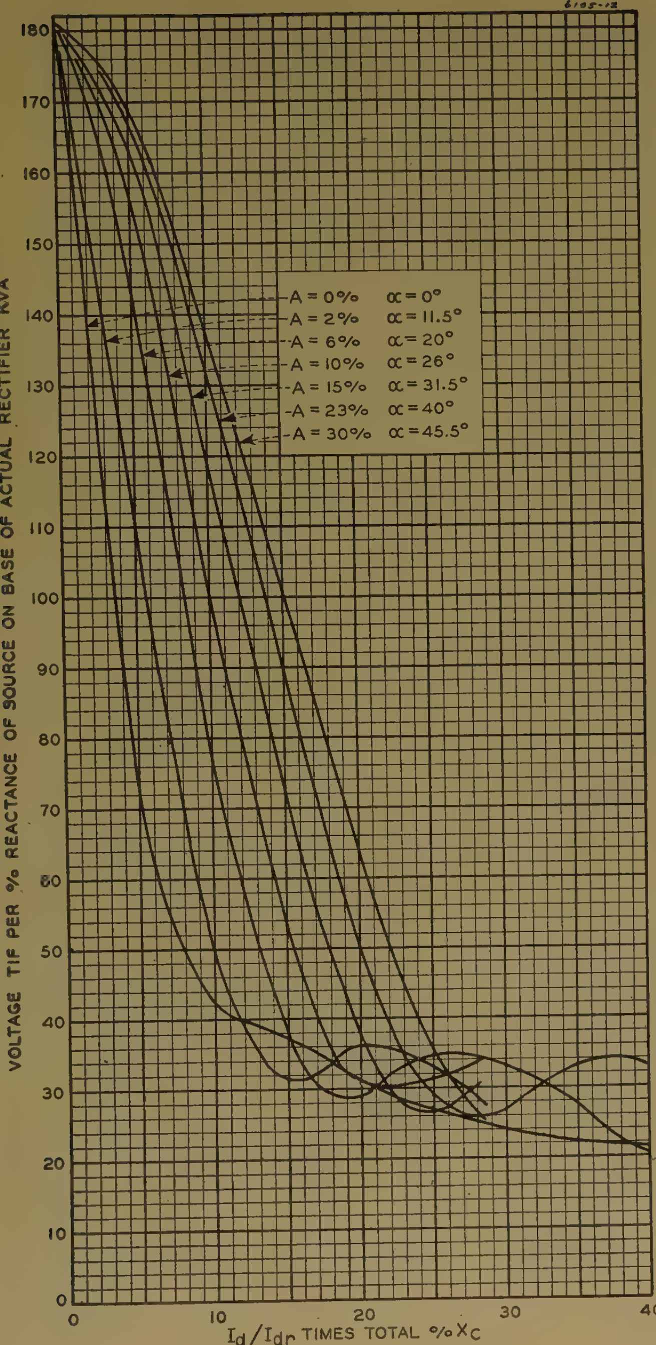


Figure 12. Six-phase rectifiers

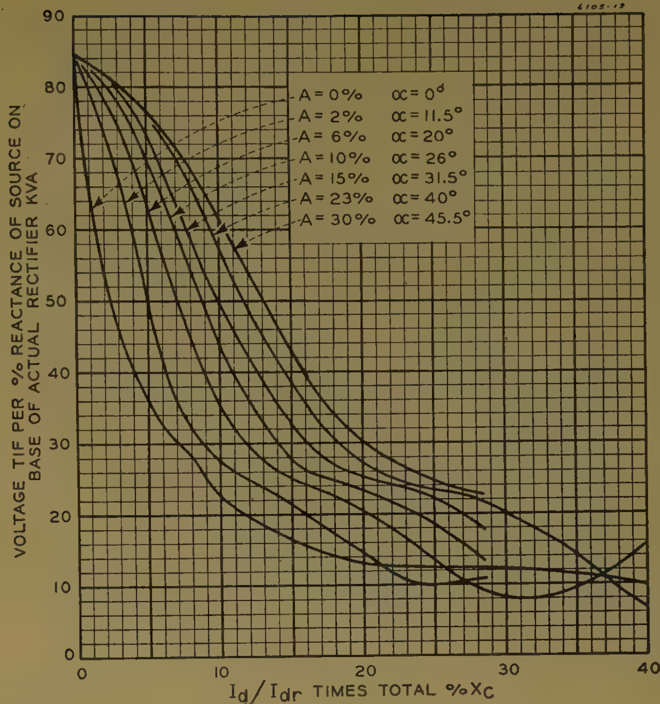


Figure 13. Twelve-phase rectifiers

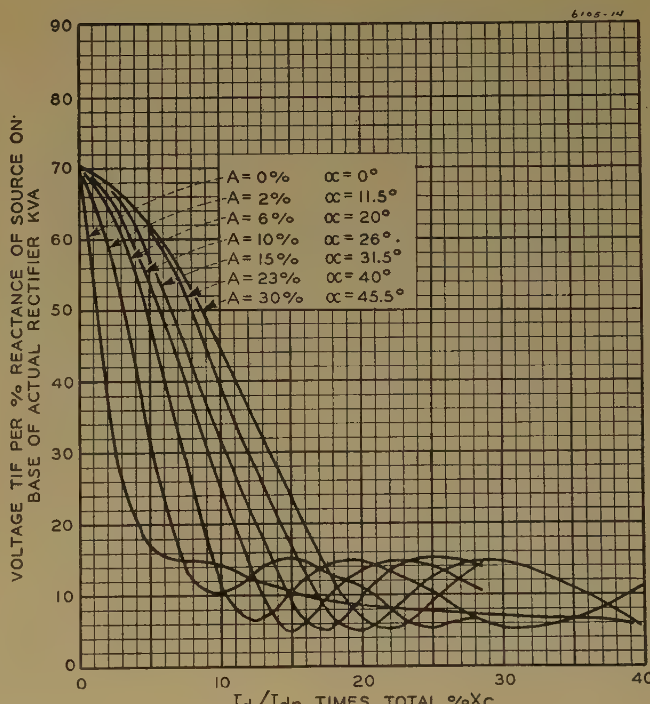


Figure 14. Twenty-four-phase rectifiers

Figures 12, 13, 14. Voltage TIF per per cent reactance of power source for 6-, 12-, and 24-phase rectifiers with different amounts of phase control

For 60-cycle systems
I_d = actual d-c load current
I_{dr} = rated d-c load current
Total $\%X_c$ = per cent commutating reactance of rectifier transformer bank and supply system on rectifier transformer rated kilovolt-ampere base
α = phase control angle of retard
A = amount of phase control or per cent voltage reduction below output voltage with zero phase control ($A = 1 - \cos\alpha$)

To find voltage TIF:

1. Determine the value of I_d/I_{dr} times total $\%X_c$ for the particular load under consideration. Where the TIF for the rated capacity only is desired, total $\%X_c$ is used directly

2. Determine the per cent power source reactance on the base of *actual* rectifier kilovolt-amperes. For partial load, this can be found by multiplying source reactance on the base of rectifier transformer rated kilovolt-amperes by the ratio I_d/I_{dr}

3. From the curve for the appropiate number of phases and amount of phase control, determine the voltage TIF per per cent source reactance

4. Multiply the value read on the curves by source reactance found in step 2. This is the estimated voltage TIF

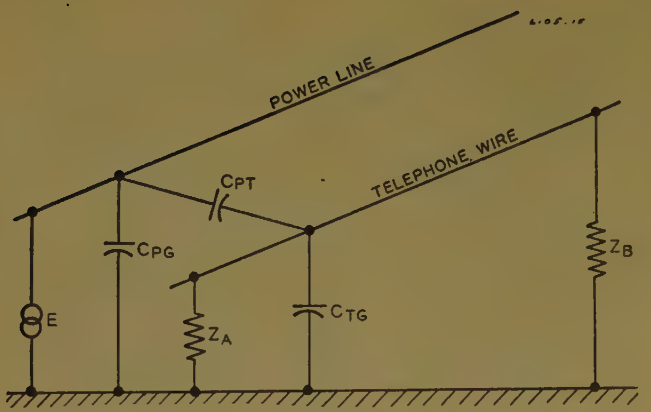


Figure 15. Nature of electric induction

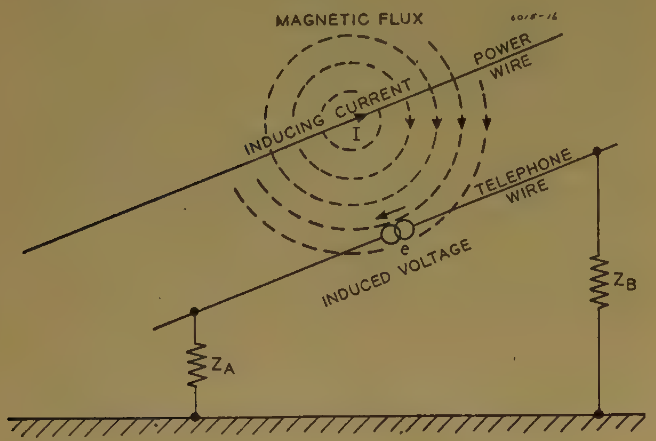


Figure 16. Nature of magnetic induction

5. From Figure 9, the I·T per kilovolt-ampere on a 1-kv circuit for total $\%X_c$ of 8.4 per cent, $\frac{I_d}{I_{dr}}$/equal to one, and phase control of ten per cent, is about 410; or on a 13-kv circuit, it is 410/13=32. The total I·T on feeder *A* is then 32×2,530=80,000.

6. From Figure 12, the voltage TIF per per cent power source reactance is about 120. The total per cent power source reactance to point *B* is 0.52+1.7=2.2 per cent. The estimated voltage TIF at point *B* is, then, 120×2.2=260.

To illustrate the method of obtaining I·T products and voltage TIF's at partial load, suppose it is desired in the example just illustrated (that is, the same as Figure 8 but with a 2,530-kilovolt-ampere rectifier transformer) to find the conditions at a *current* output of 0.6 times rated output, but with 15 per cent phase control at this load. Proceed as follows:

1. $\frac{I_d}{I_{dr}} \times \%X_c = 0.6 \times 8.4 = 5.0$ per cent.

2. I·T per kilovolt-ampere on 1-kv circuit with 15 per cent phase control is (from Figure 9) about 530. On a 13-kv circuit, it would be 530/13=41. The kilovolt-ampere input can be taken as 2,530×0.6=1,520. The total I·T on the feeder is 1,520×41=62,000.

3. The voltage TIF per per cent power source reactance is (from Figure 12) about 163. To estimate voltage TIF at point *B* in Figure 8, multiply per cent power source reactance by 0.6; that is, 0.6×2.2=1.3. Next, multiply by 163, that is, voltage TIF at *B* equals 1.3×163=212.

Figures 10, 11, 13, and 14 assume balanced operation between the 6-phase units making up the 12- and 24-phase installations; that is, it is assumed that the d-c loads are equal, that the voltages of the d-c transformer windings on all units are identical, that the differences in phase control between units are small and that the harmonic voltages on the power supply system are too small to result in unbalances. If the d-c winding voltages of the rectifier transformers differ between units (which may be caused by operation on different transformer taps) and different amounts of phase control are used, the curves no longer give accurate results, even if the d-c loads are balanced. Under such conditions, the suppression of the "normally suppressed" harmonics is likely to be impaired and partial suppression of "normally unsuppressed" harmonics may occur.

Other Forms of Conversion Equipment

Co-operative investigations have been made in connection with two installations of large frequency changers connecting 60-cycle and 25-cycle 3-phase systems.[15,39] Tests have also been made on two installations of equipment for converting from 3-phase 60 cycles to single-phase at a frequency on the order of 1,000 cycles.[38] The test results indicate that the methods outlined herein can be applied in connection with such installations without large error.

Data are not available for estimating I·T products and voltage TIF's for other forms of conversion apparatus, such as electronically controlled motors. It appears, however, that where the power supply to such devices is balanced 3-phase, estimates made along the lines discussed herein would provide at least a good starting point in studies.

The inductive effects of single-phase rectifier installations are subject to much wider variations than 3-phase installations, and the methods outlined herein cannot be applied to such installations.

Appendix I. Fundamentals of and Terms Used in Noise Frequency Co-ordination Work

In this appendix, fundamentals of and terms used in noise frequency inductive co-ordination work are discussed.

Electric and Magnetic Induction—Coupling

Voltages induced on telephone circuits by power circuit voltages are said to be caused by electric induction; voltages caused by power circuit currents are said to be the result of magnetic induction. These forms of induction exist independently and must be so treated.

The process of electric induction in its simplest form is shown in Figure 15. The power wire has on it a harmonic voltage *E* to ground.

1. Between the power wire and a nearby telephone wire, electric coupling, represented by capacitance C_{PT}, exists. The magnitude of this capacitance is directly proportional to length of exposure and increases as separation between the wires decreases. Its impedance is inversely proportional to frequency.

2. The telephone wire has finite impedances to ground, Z_A and Z_B, outside the exposure. These impedances are usually much smaller than the impedance of C_{PT}. The current to the telephone wire is therefore controlled by C_{PT}; it is directly proportional to the coupling and to frequency and magnitude of *E*. The voltage on the telephone wire is the current through C_{PT} times Z_A and Z_B in parallel; it is roughly proportional to coupling and frequency.

The magnetic induction process in its simplest form is shown in Figure 16.

1. A mutual reactance, or magnetic coupling, exists between the power wire and the nearby telephone wire by virtue of the fact that the magnetic flux associated with the power wire current links the telephone wire. The total amount of flux (per unit of power wire current) which links the telephone wire is proportional to length of exposure and increases as separation is decreased.

2. The voltage *e* induced along the telephone wire is proportional to the total flux which links it and is also proportional to frequency. The current in the telephone wire is this voltage divided by (Z_A+Z_B). The voltage to ground at one end of the exposure is this current times Z_A; at the other, times Z_B.

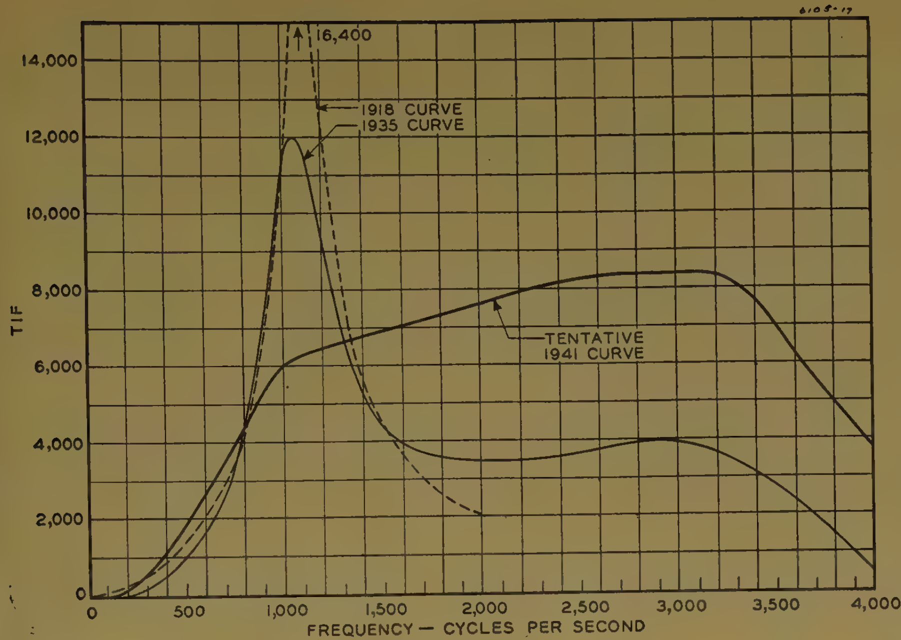


Figure 17. TIF weightings

Voltage TIF and I·T Products

The frequencies in human speech range from the order of 125 cycles to upwards of 10,000 cycles per second. A much narrower band of frequencies is sufficient to carry on a satisfactory conversation and most telephone circuits are designed to transmit such a narrower band.

Each harmonic voltage or current in a power system induces a voltage of the same frequency in a paralleling telephone circuit which, in turn, may actuate a telephone receiver and cause an audible sound. The composite of the sounds of all such frequencies present usually results in a humming sound commonly called noise. Different frequencies in this noise have different interfering effects[27,29] depending on the characteristics of the telephone circuits, receiver, the human ear, and other factors; relative interfering effects are called noise weightings and have been ascertained by extensive tests. The composite of the noise voltages in the circuit can be measured by means of a noise measuring set which contains networks having frequency-transmission characteristics in accordance with the noise weightings, and a meter which adds components as the square root of the sum of the squares.

Since the effects of both electric and magnetic coupling are directly proportional to frequency, the relative noise influence of voltages and currents on the power system is proportional to the product of noise weighting and frequency. For any frequency, this product (times a constant) is known as TIF (telephone influence factor) weighting. TIF weighting curves as functions of frequency are shown in Figure 17.

There are three different sets of TIF weightings which reflect changes in transmission-frequency characteristics of telephone circuits and instruments over a long period, namely:

1. The "1918" weightings. These were known as telephone interference factor weightings, and this name is still used to distinguish them from the later weightings which are called telephone influence factor weightings.

2. The "1935" weightings.[28]

3. The tentative "1941" weightings. These weightings have not been standardized yet.

The very small weightings of 60 cycles explains why, from the noise frequency induction standpoint, only the harmonics are of interest.

While much induction work is carried on on the basis of individual harmonics, it frequently is desirable to use an over-all measure of the influence of a current or a voltage in a power circuit. This over-all measure is obtained by multiplying the magnitude of each harmonic present (amperes, for current; kilovolts for voltage) by its TIF weighting, and taking the square root of the sum of the squares of these products. The result is, for current, I·T product; for voltage, KV·T product. These products can be measured directly by the use of a noise measuring set and a TIF coupler, the latter having a transmission-frequency characteristic directly proportional to frequency. So-called TIF meters also can be used.

Balanced, Residual, and Ground Return Currents and Voltages

Suppose in Figure 16 there were another power wire very close to the one shown there and that it carried a current equal and opposite to the current in the original wire; such a set of currents would be called balanced. The flux cutting the telephone wire then would be only the difference between the fluxes at that point arising from the currents in the two power wires. If the separation between the power wires were small, this difference would be small and hence the coupling would be small as compared to that for current which goes out on one or more wires and returns through the earth, that is, ground return current.

A similar analysis with respect to voltages would indicate the same conclusions.

For a power circuit of any number of phases and wires, the following terms apply:

1. Ground return current is the vector sum of the

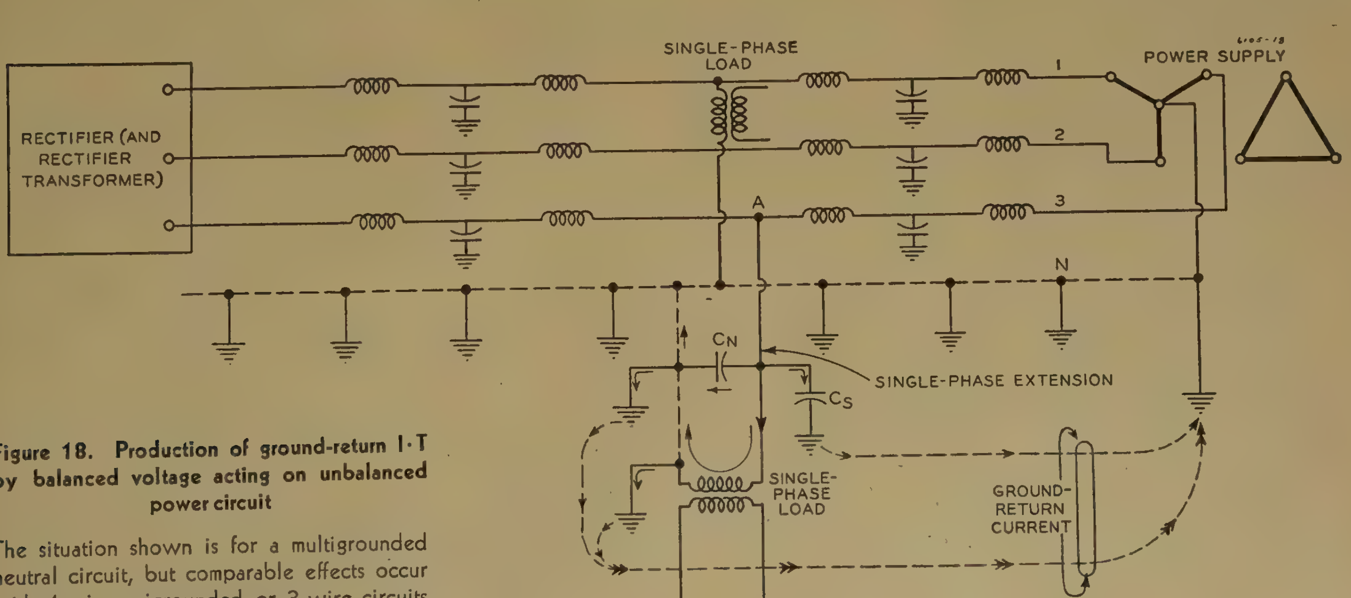


Figure 18. Production of ground-return I·T by balanced voltage acting on unbalanced power circuit

The situation shown is for a multigrounded neutral circuit, but comparable effects occur with 4-wire unigrounded or 3-wire circuits

actual currents in all the wires of the circuit.

2. Residual current is the vector sum of the actual currents in the phase wires. In a 3-phase 3-wire circuit, residual and ground return currents are the same. In a 4-wire 3-phase circuit, part of this residual current may stay in the neutral wire, and the influence of this part is about the same as if it were balanced. In a 2-wire circuit consisting of an extension of a phase wire and neutral of a 3-phase 4-wire circuit, phase and residual current are the same; part of this residual current is ground return.

3. Balanced currents are those components of the actual phase currents whose vector sum is zero.

4. Residual voltage is the vector sum of the voltages between phase wires and earth.

5. Balanced voltages are those components of the actual phase-to-ground voltages whose vector sum is zero.

The over-all influence of any one of these components is measured by its I·T or KV·T product. For distribution circuits, and in many cases for transmission circuits, over-all influence is controlled by ground return I·T, or residual KV·T. In dealing with voltages of known magnitude, voltage TIF is frequently used instead of KV·T; it is KV·T divided by voltage in kilovolts. For residual voltages on 3-phase generators, residual component voltage TIF is sometimes used; it is one third residual KV·T divided by phase-to-neutral voltage.

In a 3-phase circuit with no single-phase extensions or loads, the ground return I·T and the residual KV·T are usually small except for triple harmonics (third and odd multiples thereof). If, however, there are single-phase taps or loads, ground return harmonic currents can occur even though the harmonic voltages are impressed in a perfectly balanced manner, as by a rectifier. The conditions leading to this effect are shown in Figure 18. Here is shown a feeder from a power source to a rectifier; there is a single-phase extension. For simplicity in explanation, a multigrounded neutral circuit is shown; comparable effects may occur with ungrounded neutral or 3-wire circuits as explained later.

As explained in "Nature of Problem" in the main body of this report, the harmonic currents produced by the rectifier load flow through the phase wire reactances and produce harmonic voltages. The harmonic voltage-to-neutral thus caused at point *A* in Figure 18 is impressed on the single-phase tap. Capacitance exists between the phase wire in the tap and the neutral (C_N) and between the phase wire and ground (C_S). The load likewise presents a finite impedance between the phase wire and the neutral and ground in parallel. Harmonic currents flow through the capacitance and load impedances. Part of this current flows in the neutral wire (about 40 per cent in the usual case) and part in the ground (usually about 60 per cent). The ground current returns to the power circuit via the neutral ground on the supply transformer bank. Thus on the tap itself and on the portion of the 3-phase circuit between the tap and the power source, there is ground return harmonic current. Harmonic currents through single-phase loads connected directly to the 3-phase line produce ground return harmonic current in a similar fashion. In an actual circuit, there are usually numerous single-phase taps and single-phase loads; ground return current can circulate between one tap or load and other taps or loads, so that the ground return I·T is likely to be different at different places along a 3-phase circuit.

That comparable effects can occur on long unigrounded neutral or 3-wire circuits can be shown as follows:

1. Consider first a 4-wire unigrounded neutral circuit with a single-phase tap. Imagine that, in Figure 18, all grounds on the neutral are removed except the one at the power supply bank and series inductances and shunt capacitance are distributed along the neutral as well as the phase wires. Harmonic current can still flow through the capacitance between the phase wire and ground; this current is ground return. Furthermore, under some conditions, substantial voltages to ground can exist on the neutral at harmonic frequencies; they may cause harmonic currents through the neutral-to-ground capacitance, and these currents are ground return.

2. On a 3-wire circuit, the single-phase tap would be connected to two of the phase wires. Harmonic currents can flow through the capacitances to ground from both phase wires and become ground return. Even if there is no neutral-to-ground connection at the power supply bank, these ground return currents can exist; they return to the system through the capacitance to ground of the remaining phase wire in other parts of the system.

As a general rule, overhead urban distribution systems are so short that the capacitances can be neglected. Ground return current can occur in such systems only if the neutral is multigrounded. Rural systems, however, are usually long enough so that the harmonic currents through the capacitance are frequently controlling; substantial ground return I·T can occur regardless of whether or not a neutral wire is present or where or how the system is grounded.

There are three other methods by which ground return harmonic current can be caused by balanced impressed harmonic voltages or currents (as by a rectifier), namely:

1. The currents in the phase wires magnetically induce voltages along the neutral or ground wire. If the phase wires are not disposed symmetrically with respect to the neutral, the voltages induced along the neutral wire by the balanced currents in the three individual phase wires will not be equal. The vector sum can, in a multigrounded neutral circuit or a transmission circuit equipped with a ground wire, cause current to circulate between the neutral or ground wire and ground; this current is ground return. This effect usually is important only in cases where the circuit is very well balanced otherwise.

2. On a multigrounded neutral urban circuit with capacitors, small differences between the reactances of the individual phase wires can cause substantial ground return harmonic current if the capacitors and circuit are very close to resonance at one of the harmonic voltages present. This phenomenon is discussed more fully in Appendix II.

3. In some cases the minor differences in the reactances of the phase wires can cause ground return harmonic currents even in the absence of capacitors. Such ground return currents can occur only when

(*a*). The circuit is long enough to permit appreciable currents through the natural capacitances, *or*

(*b*). There are two or more transformer bank neutral-to-ground connections between which the ground return current can circulate.

Metallic and Longitudinal Telephone Voltages and Currents—Susceptiveness

Assume that in Figure 16 another telephone wire is added parallel and close to the first. The second wire will be linked by nearly the same flux as the first, and the difference in the voltages along the two wires will be a small fraction of that along either one. If these two wires are used as a metallic telephone circuit, the noise voltage acting around this circuit will be a small proportion of that which would act in a telephone circuit consisting of one wire with ground return, that is, a ground return circuit.

The same analysis could be carried through for electric induction with the same conclusions.

Thus in a metallic telephone circuit, there is longitudinal circuit induction which is the voltage along or to ground on the conductors in parallel, and direct metallic circuit induction, which is the difference between the voltages along or to ground on the two wires, the direct metallic circuit induction being much the smaller.

Direct metallic circuit induction can be reduced still further by transposing the telephone circuit, that is, interchanging the positions of the conductors periodically so that the voltages on both tend to be equal. In open wire circuits the reduction obtainable, while large, is limited by unavoidable irregularities in exposure conditions. In cable circuits the pairs are twisted so as to be continuously transposed, and direct metallic circuit induction can be neglected. In addition, the telephone cable sheath provides practically perfect shielding against electric induction if grounded at one or more points, and it provides some shielding against magnetic induction if grounded at both ends, the degree of shielding being limited by the resistance of the sheath and of the grounds.

Longitudinal circuit induction can cause noise in a metallic circuit by acting on tele-

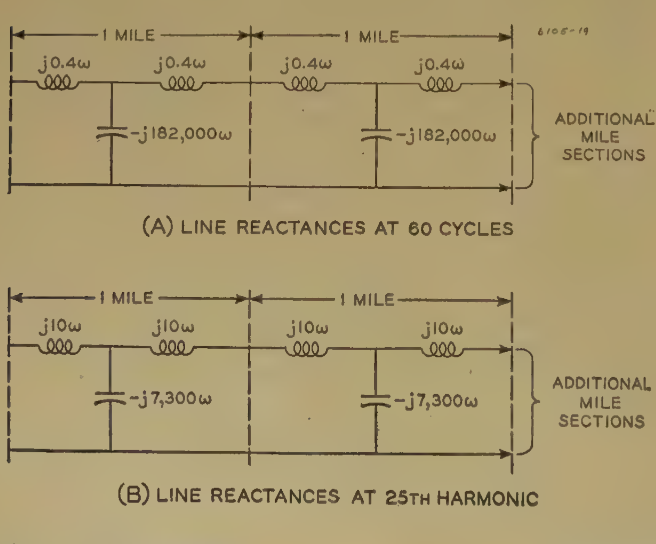


Figure 19. Phase-to-neutral inductive and capacitive reactances of typical transmission line

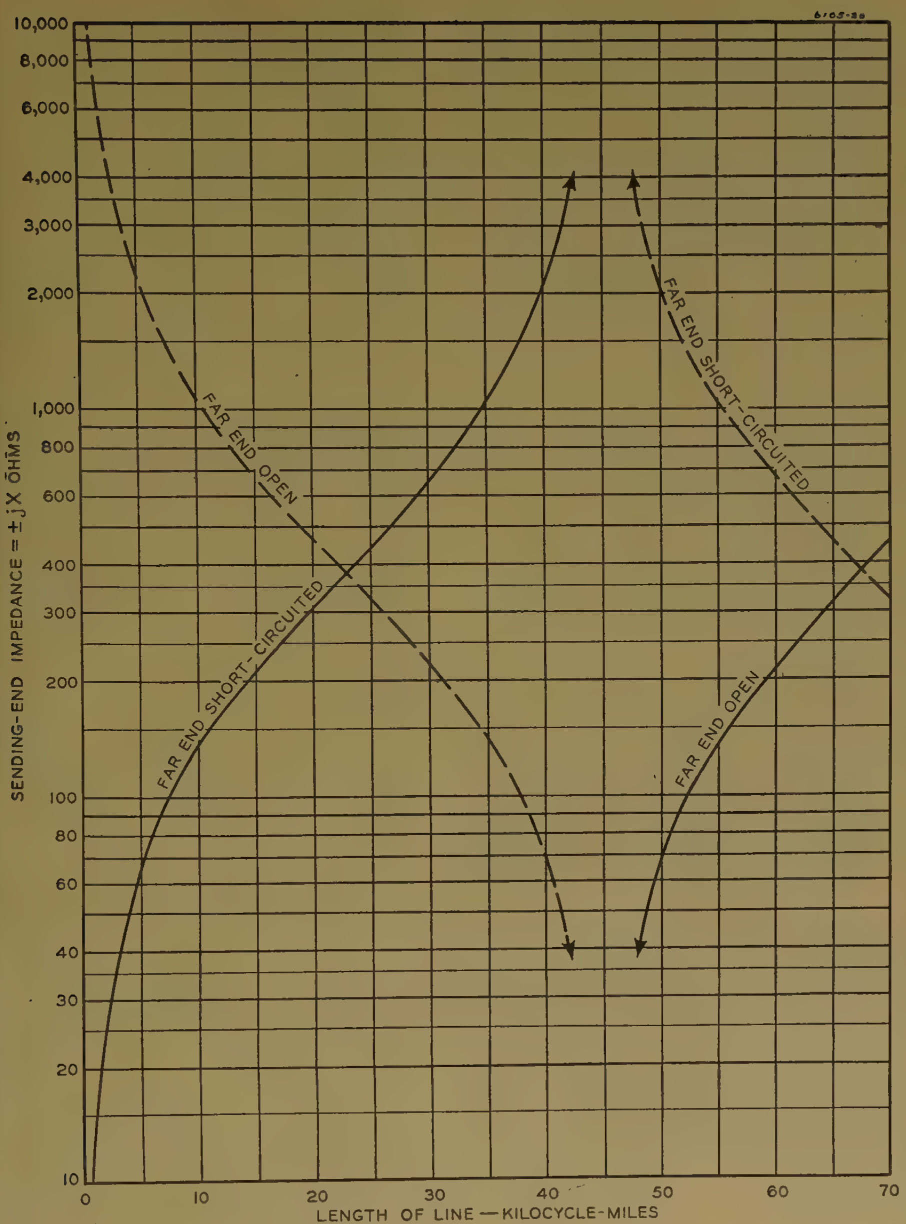


Figure 20. Sending-end phase-to-neutral impedance of typical transmission line with far end open and short-circuited

——— Sending-end inductive reactance $=+jX$

– – – – – Sending-end capacitive reactance $=-jX$

Impedances are for condition of no losses in circuit. For effect of losses, see text

phone circuit unbalances. These unbalances are of two types:

1. Series impedance unbalances, that is, differences in the impedances of the two wires of the circuit.

2. Shunt admittance unbalances, that is, differences in the admittances between the two wires of the circuit and ground.

Unbalances can occur in line, station, or central office equipment. Typical types of unbalances in these three locations are:

1. Lines. Series unbalances may be caused by imperfect joints or by variations in cross section or composition of the two conductors of a circuit. Shunt unbalances may be caused by such things as missing insulators, foliage touching the wires, transposition irregularities, and unbalances in capacitance to ground in cables.

2. Stations. Shunt unbalances only need to be considered. The principal unbalance is the ringer-to-ground connection used for party-line ringing. Its magnitude depends on the impedance of the ringer and the details of the subset connections.

3. Central office. Series unbalances can occur on account of unbalances in relays used for supervision, differences in capacitance of condensers used to block direct current paths, and so forth. Shunt unbalances may be caused by differences in the inductances of relay, retard or repeating coil windings connected between the wires and ground, differences in the capacitances to ground of wires, repeating coil windings, and other factors.

The degree to which a telephone circuit can be affected by given extraneous electric and magnetic fields is called its susceptiveness. Susceptiveness depends on such things as balance and adequacy of transpositions, and on the frequency response of the circuits and instruments and the level (magnitude) of the telephonic current and voltages at the point where the exposure exists.

Inductive Co-ordination

It is theoretically possible to have power circuits with practically zero influence (for example, by having no harmonics), they would not cause noise in any telephone circuit regardless of its susceptiveness. It is theoretically possible to have telephone circuits of practically zero susceptiveness (for example, by having perfect balance and transpositions); they would not be noisy regardless of the influence of nearby power circuits. Also, of course, if the power and telephone systems were separated by large distances, the coupling, and hence the noise, would be negligible. None of these methods is practicable, in general, in the present state of the art. The practical job of noise frequency inductive co-ordination consists of so controlling coupling, influence, and susceptiveness within practical limits that telephone circuit noise is controlled adequately at minimum cost and maximum satisfaction to both power and telephone customers.

Appendix II. Propagation of Harmonic Currents and Voltages

Line Characteristics

For short feeders between a source of power supply (generator or substation) and a rectifier installation, only the inductive reactances need be considered in estimating harmonic voltages and currents. But for longer circuits such as transmission or rural distribution circuits, the capacitances between the different wires and between the wires and ground must be taken into account when dealing with harmonic frequencies.

Why the capacitances associated with the wires have a greater effect at harmonic frequencies than at 60 cycles can be shown by considering the constants of a typical transmission line as illustrated in Figure 19. Here are shown the series inductive reactances and shunt capacitive reactances on a phase-to-neutral basis for such a line (Figure 19A) at 60 cycles and (Figure 19B) at the twenty-fifth harmonic. It can be seen that at 60 cycles the shunt capacitive reactances are very large compared with the series inductive reactances; a line of short or moderate length can be considered simply as consisting of $j0.8$ ohm per mile. But at the twenty-fifth harmonic, the series impedances are increased 25 to 1, and the shunt impedances are decreased 25 to 1. At the twenty-fifth harmonic, the total series impedance becomes as large as the total shunt impedance in 19.1 miles of line (that is, $19.1\times20=382=7{,}300/19.1$), whereas in the case of 60 cycles, this equality does not occur until the line is about 480 miles long.

To obtain an idea of how, quantitatively, this difference affects line performance, consider Figure 20, which shows the sending-end impedance of a typical transmission line with the far end open and with it short-circuited, as a function of length and frequency, neglecting losses. These curves are computed from the formulas:

Sending-end impedance with far end open $=-jZ_0\cot\theta$

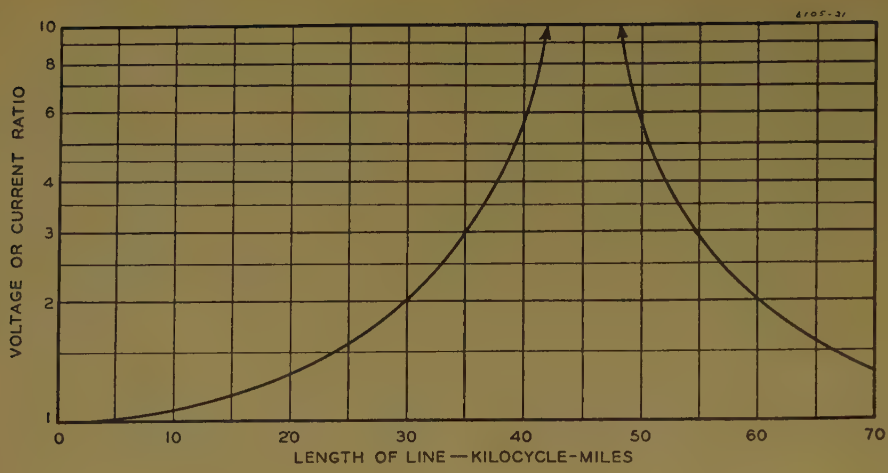


Figure 21. Ratio of current at far end to current at near end with far end short-circuited and ratio of voltage at far end to voltage at near end with far end open

Ratios are for condition of no losses in circuit. For effect of losses, see text

Sending-end impedance with far end short-circuited $= jZ_0 \tan \theta$

where

Z_0 = characteristic impedance to neutral (380 ohms in the example)

θ = line angle (or phase shift) in degrees = frequency times length times 360 divided by velocity of propagation

Since, in a smooth circuit without losses (a good approximation for power transmission circuits at harmonic frequencies), the velocity is independent of frequency, the curves can be plotted using length times frequency, that is, kilocycle-miles, as abscissas.* For example, a 100-mile line at 60 cycles is 0.06 kilocycle times 100 miles = 6 kilocycle-miles; a 20-mile line at 1,500 cycles is 30 kilocycle-miles, and so forth. For such lines the velocity is about 180,000 miles per second.

With a velocity of 180,000 miles per second, θ, or phase shift, is two degrees per mile per kilocycle, which can be derived as follows:

$$\text{Wave length at frequency } f = \frac{180{,}000 \text{ miles}}{f}$$

Wave length = 360 degrees

$$\text{Whence } \frac{180{,}000}{f} \text{ miles} = 360 \text{ degrees}$$

$$\text{Degrees per mile} = \frac{360 f}{180{,}000} = 2° \times \frac{f}{1{,}000}$$

If f is in kilocycles, degrees per mile = two degrees times f (in kilocycles).

It will be noted from Figure 20 that

1. With the far end open, the impedance looking into the near end is capacitive reactance up to 45 kilocycle-miles; with the far end short-circuited, it is inductive reactance up to that length. Above 45 kilocycle-miles, the signs are reversed.

2. At 22.5 and 67.5 kilocycle-miles, the impedances with the far end open and short-circuited are the same in magnitude but opposite in sign.

3. At 45 kilocycle-miles, the impedance with the far end open is very low; with the far end short-circuited, it is very high. That is, at this length a short circuit on the far end tends to look like an open circuit at the near end; an open circuit at the far end tends to look like a short circuit at the near end. Actually, the near end impedances would go to zero or infinity except for the losses in the circuit. The significance of 45 kilocycle-miles is that it is "quarter wave length." That is, 45×2 degrees = 90 degrees, which is one quarter of 360 degrees. The effects at this length are sometimes referred to as "natural line resonance" because a line of this length with the far end open looks from the sending end like a series resonant circuit; with the far end short-circuited, it looks from the sending end like a parallel resonant circuit.

* That frequency and length can be combined is also obvious from the fact that the inductive reactance as well as the capacitive admittance (reciprocal of reactance) are directly proportional to length and frequency.

Another factor of importance in considering propagation is what happens to harmonic currents and voltages as they traverse the line from the sending-end to the far end. Figure 21 shows, for the same line as used in Figures 19 and 20, the ratio of voltage at the far end to that at the near end with the far end open, and the ratio of current at the far end to that at the near end with the far end short-circuited. This curve is based on the formula

$$R = \frac{1}{\cos \theta}$$

where

R = ratio of voltage at far end to that at near end with far end open, or ratio of current at far end to that at near end with far end short-circuited.

θ = line angle in degrees = two degrees times kilocycle-miles.

The abscissas are again in terms of kilocycle-miles. It is of some interest to note that these two ratios are shown by the same curve, that is, the voltage ratio with the far end open is the same as the current ratio with the far end short-circuited. The quarter wave length effect is again evident.

Here again the difference between the way a line performs at 60 cycles and at harmonic frequency may be noted. Consider a line 50 miles long with the far end open. The data for a few typical frequencies are set down in Table XI.

In most practical cases the far end of a line is not open nor short-circuited. While the treatment of the general case is beyond the scope of this appendix, it may be said that for transmission circuits the impedance at harmonic frequencies across the far end is frequently almost pure inductive or capacitive reactance. In such cases the effects shown in Figures 20 and 21 occur, but for different line lengths than shown there.

Equipment

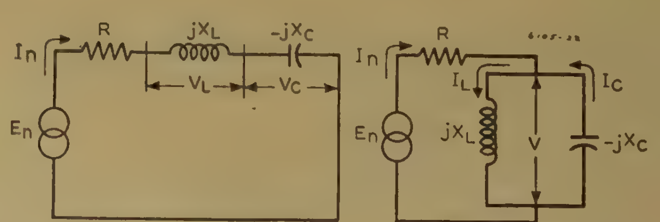
Transformers and rotating machinery present almost pure inductive reactances; the impedance at harmonic n is very closely equal to n times the 60-cycle impedance

Table XI

Harmonic	Frequency	Kilocycle-Miles	Ratio of Far End Volts to Near End Volts (From Figure 21)	Sending-End Impedance (From Figure 20)
1	60	3	1.006	$-j3{,}650$
7	420	21	1.35	$-j425$
17	1,020	51	4.8	$+j84$
25	1,500	75	1.15*	$+j660$**

* Obtained from fact that the ratio for L kilocycle-miles is the same as the ratio for (90-L) kilocycle-miles, that is, the ratio for 75 kilocycle-miles is the same as the ratio for 15 kilocycle-miles.

** Obtained from the fact that sending-end impedance with far end open of a line L kilocycle-miles long is the same as sending-end impedance with far end short-circuited of line (L-45) kilocycle-miles long, that is, sending-end impedance with far end open of line 75 kilocycle-miles long is the same as impedance with far end short-circuited of line 30 kilocycle-miles long.



(A) SERIES RESONANCE (B) PARALLEL RESONANCE

Figure 22. Series and parallel resonance

For series resonance,

$$I_n = \frac{E_n}{R + jX_L - jX_c}$$

If $X_L = X_c$, then

$I_n = E_n/R$

$V_L = I_n X_L$

and

$V_c = I_n X_c$

For parallel resonance,

$$I_n = \frac{E_n}{R + \dfrac{(jX_L)(-jX_c)}{jX_L - jX_c}}$$

If $X_L = X_c$, then

$$I_n = \frac{E_n}{R + \dfrac{X_L X_c}{0}} = 0$$

$V = E_n$

$I_L = V/jX_L = E_n/jX_L$

and

$I_c = V/-jX_c = E_n/-jX_c$

A capacitor presents practically pure capacitive reactance of a magnitude equal to its 60-cycle reactance divided by n. For generators and synchronous motors, the average of the negative sequence and subtransient 60-cycle reactances is used in estimating harmonic reactance.

Transformers step up or step down impedances by the square of the turn ratio; that is, if an impedance Z is connected to the low side of a transformer having a voltage ratio K, it appears when looking through the transformer from the high side as ZK^2. Where transformers are used, the impedances of equipment on one side can be transferred to the other either by computing them as if they were actually operated on the desired voltage base or by computing them on their own voltage base and multiplying by the square of the transformer ratio.

Resonance Between Equipment and Lines

It has been noted in Figure 20 that over comparatively wide ranges of frequency times length, the sending-end impedance of lines is capacitive reactance. Since the reactance of transformers and rotating machinery is inductive, it is not uncommon to find a frequency at which the reactances of the apparatus and the line are equal and opposite.

When equal and opposite reactances are connected together, a condition known as "resonance" occurs. Two variations of resonance can be recognized, that is, series resonance and parallel resonance, as illustrated in Figure 22. There it is shown that:

1. For series resonance, the current through and the voltages across the inductive and capacitive reactances are limited only by the resistance in the circuit. Thus, in a low loss series resonant circuit, large voltages and currents can occur even with relatively low impressed voltage.

2. For parallel resonance, the current in the reactive elements is equal to the impressed voltage divided by the reactance of the element. The voltage across the circuit does not exceed the internal voltage of the source. Consequently, large currents can occur only when the internal voltage of the source is high or the reactances of the elements are low. On the other hand, the current through and the voltage across the reactive elements are relatively unaffected by the internal resistance of the source.

In Figure 23 are shown examples of possible resonance conditions which may be set up at a harmonic frequency between the inductive reactance of equipment and the capacitive sending-end reactance of lines. It is assumed that the source is putting out a harmonic of the resonant frequency.

Rotating machinery and transformers usually have low internal resistance so that substantial increases in harmonic voltages and currents at particular frequencies can be caused by either of the types of situation illustrated. Transmission lines with harmonics put on at remote points usually also act as sources with low internal resistance when considering harmonics impressed on distribution systems. This is because

1. The impedance looking into a transmission line, to which most of the apparatus connected has inductive reactance, usually has a relatively small resistive component.

2. Further, the resistive component (as well as the reactive component) is stepped down by the square of the transformer ratio when considered on the distribution circuit voltage base.

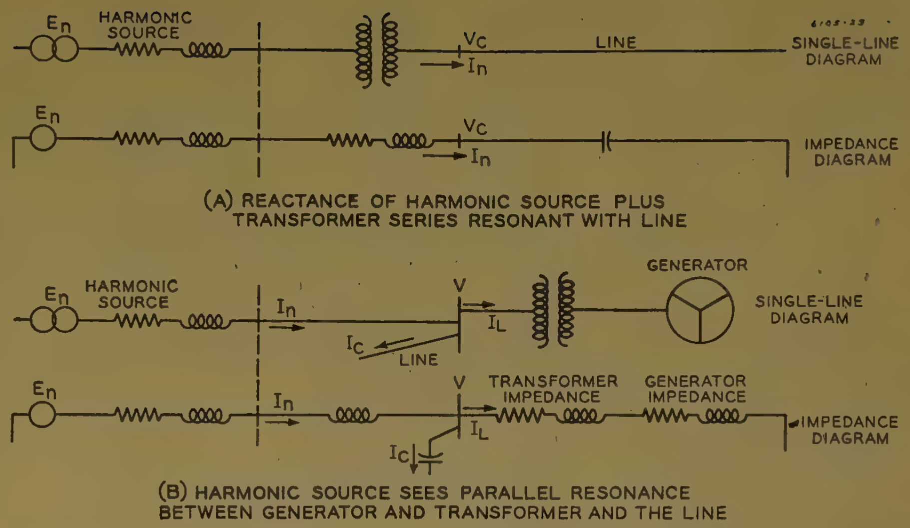


Figure 23. Examples of resonance between equipment and lines

On the other hand, the internal resistance of a rectifier, considered as a source of harmonics, is significant; the resistance is an inverse function of the load. Therefore, except for the larger rectifier installations, which are characterized by low internal resistance, large increases in harmonic currents or voltages at the sending ends of lines because of resonance are likely to occur only in cases of parallel resonance (as in Figure 23B).

Fortunately, most of the resonances encountered in practice are relatively sharp, and there is usually no harmonic present of exactly the resonant frequency. Consequently, the increases in harmonic voltage and current because of resonance are only moderate in most cases, although in rare cases these increases may be ten or more to one.

The ratios of harmonic currents and voltages at the two ends of a line (shown in Figure 21) are independent of whether the magnitudes at the sending end are affected by resonance.

Rural Distribution Circuits

Propagation of harmonics over rural distribution circuits differs from that over transmission circuits in two important respects:

1. The impedance presented by load is largely resistive in series with inductive reactance, the latter being caused by the load transformers. At night, and sometimes during the day, the load is light and the line operates essentially with the far end open.

2. A large proportion of the line mileage is likely to be single-phase, so that the ratio of ground-return harmonic current to phase harmonic current is much higher than for a transmission line. For single-phase multigrounded neutral rural lines, the ratio of ground return to phase harmonic current is on the order of 0.6; for ungrounded neutral or delta lines, the ratio is variable and it may be as high as unity.

Because of the facts that over substantial periods, rural power lines act as if they were open and that the line length is moderate, the sending-end impedance is likely to be capacitive reactance at the frequencies of the important harmonics. The sizes of transformers used to supply rural lines are frequently such that resonance occurs somewhere between the fifth and twenty-fifth harmonics. During heavier load periods, this resonance is considerably damped by the load; as a result, the noise influence of a rural power circuit usually is substantially higher at night than during the periods when the load is heavier.

In dealing with the single-phase branches of distribution circuits, it is rather common to use impedance-to-ground instead of impedance-to-neutral. Impedance-to-ground is phase-to-ground voltage divided by ground-return current or, conversely, ground-return current is phase-to-ground voltage divided by impedance-to-ground. Figures 20 and 21 can be used to obtain an idea of what happens on a single-phase rural line with the far end open or short-circuited when a harmonic voltage is impressed on it by making the following changes:

1. Before using the curves, multiply the actual kilocycle-miles by the following factors to take account of the lower velocity of propagation:

Ungrounded neutral	1.3
Multigrounded neutral	1.1

2. Multiply the ordinates in Figure 20 by about two. The ordinates in Figure 21 are unchanged.

There are many factors affecting the influence of a rural power line which cannot be discussed here. References 11, 12, and 23 discuss these matters in detail.

Urban Distribution Systems

Overhead urban distribution systems are usually so short that the effect of capacitance can be neglected except possibly for unusually long lightly loaded circuits with unusually high impressed magnitudes of the higher frequency harmonics. Consequently, in the absence of capacitors, the only path for harmonic currents is through the loads. Unlike rural distribution circuits, ground-return harmonic current can flow only where the neutral is multigrounded.

Where capacitors are used, they frequently present lower impedances at harmonic frequencies than do the loads. It not infrequently occurs that the resonant frequency of the capacitors and supply transformers and line is between the fifth and about the eleventh harmonic. In some cases this results in relatively large harmonic currents and voltages at one of these frequencies. In some cases the small differences between series reactances of the three

Table XII. Summary of Studies Made in a Few Selected Cases of Rectifier Installations

Case	Size, Kilowatts	Number of Phases	Feeder A of Figure 5	Part B of Figure 5	Part C of Figure 5	Supply Transformer D, Figure 5	Supply System N, Figure 5	I·T* In A	I·T* In N	Voltage TIF* On B	Voltage TIF* On C	Co-ordination Problems	Remarks
A	21,600 +5,400	12 6	—	—	—	—	Extensive 154 and 120 kv	—	32,000 on 120 kv	—	—	Extensive over several states	Changed to 30-phase with condition as indicated in example A[1]
A[1]	27,000	30	—	—	—	—	Extensive 154 and 120 kv	—	2,600 on 120 kv	—	—	None	—
B	41,000 +20,500	48 24	—	13-kv feeding distribution	—	75,000 kva	Extensive 115 and 230 kv	—	11,000 on 115 kv	50(E)	—	Moderately extensive without capacitors connected	Large capacitor bank for power purposes reduced effect to negligible value
C	114,800	24 36 48	—	—	—	—	Extensive 230, 69, and 115 kv	—	11,000 on 230 kv	—	—	Not very extensive because of particular layout of power system. Potentially important if power interconnections changed	Tentative plans made to change 24-phase to 48-phase. Action depends on future of plant
D	32,500	36	—	—	—	36,000 kva	110 kv	—	5,000	—	—	Severe caused primarily by quarter wave length power line at 1,020 cycles, particularly when one 6-phase rectifier unit shut down	Change in power layout and removal of minor irregularities in the rectifier phase shifting arrangements corrected condition
E	2,000	6	13-kv cable direct to generator bus	Urban distribution	—	—	Generating station, 13- and 26-kv cables feeding urban and suburban distribution	42,000	42,000	60	20**	A few	Resonance between cables and generator at 1,020 and 1,140 cycles resulted in substantial increase in influence. Change to 12-phase operation practically eliminated effect
F	1,000	12	13 kv. overhead	Urban distribution	None	None	Metropolitan cable and open-wire network	17,000	17,000	85–125 Value depends on method of feed	—	Fairly extensive to parts *A* and *B* only	Filter installed which reduced voltage TIF and I·T about 4.5 to 1 and eliminated problem
G	1,000	12	13 kv, overhead	None	None	None	Metropolitan cable and open-wire network	19,000	19,000	—	—	None	No problem because of absence of parts *B* and *C*
H	3,000	6	12-kv cable	—	—	—	Extensive urban cable network	60,000(E)	60,000(E)	—	—	No a-c problem. Filter required on d-c side	Substantial exposure of telephone cable circuits to d-c feeders and trolleys resulted in telephone noise problem
I	200	6	8 miles, 13 kv, semirural	40 miles, 13 kv, semirural	—	—	Extensive high-voltage network	2,300	Negligible	100(E)	—	Extensive because of part *B*	Filter installed; reduced voltage TIF and I·T about 5 to 1. Power factor improvement provided by filter turned out to be valuable
J	1,300	6	—	—	—	—	Fairly extensive 33 kv	—	5,000	—	—	Severe over about 600 square miles	Filter applied which eliminated problem
K	600	6	6.9 kv, industrial	6.9 kv, industrial and suburban	6.9 kv, industrial and suburban	—	25-kv heavy industrial	24,000–16,000(E)***	6,000–4,000(E)***	120–335***	20–50***	Little or none	Appears to be some likelihood of problems if power or telephone conditions change
L	400	6	6.9 kv, semirural	6.9 kv, semirural	6.9 kv, semirural	—	—	16,000(E)	—	190	—	Severe locally	Filter installed; reduced TIF and I·T about 4 to 1

* Measured except where noted by (E)—estimated.

** Voltage TIF at generator bus.

*** Range among 4 similar installations.

phase-wires (because of mutual inductance between them or between them and a multi-grounded neutral) may cause a fairly large proportion of this harmonic current to become ground return. This can be illustrated by the following (rather extreme) example:

1. Assume a 180-kva capacitor bank located two miles away on a 4-kv feeder from a 3,000-kva supply transformer bank with five per cent reactance. Assume 60-cycle phase-to-neutral impedance as follows:

Transformers	$j0.29$	ohm
Line:		
Phase 1	$j1.6$	ohms
Phase 2	$j1.65$	ohms
Phase 3	$j1.71$	ohms
Capacitors	$-j96.$	ohms

2. Assume that a seventh harmonic voltage exists in the circuit impressed with the power supply. The reactances at that frequency are:

Transformers	$j2.0$	ohms
Line:		
Phase 1	$j11.2$	ohms
Phase 2	$j11.6$	ohms
Phase 3	$j12.0$	ohms
Capacitors	$-j13.7$	ohms

Totals are:

Phase 1 $= j13.2 - j13.7 = -j0.5$ ohm
Phase 2 $= j13.6 - j13.7 = -j0.1$ ohm
Phase 3 $= j14.0 - j13.7 = \ j0.3$ ohm

3. If the harmonic voltages impressed on the three phases are equal, it can be seen that (neglecting resistances) the harmonic current in phase 2 will be about five times that in phase 1 and three times that in phase 3. Also, the current in phases 1 and 2 lead the impressed voltages; in phase 3 it lags the impressed voltage. The vector sum will be larger than the current in any of the phases; about 60 per cent of this vector sum will be ground return.

Examples From Field Experience

A few examples of situations encountered in the field which illustrate some of the foregoing principles are outlined:

1. One of the early large electrometallurgical rectifier installations was operated for a time with several 12-phase units and one 6-phase unit. The voltage TIF on the 154-kv supply at the rectifier was about 75. After traversing over 150 miles of transmission line, passing by a large generating station, and being impressed on a rural line, the voltage TIF was close to 500. The voltage TIF at numerous places within about 200 miles was higher than at the rectifier.

2. There have been cases where difficulties were encountered in the operation of rectifiers because of excessive fifth harmonic voltage either because of high impedance at that frequency looking out from the rectifier because of parallel resonance or because of stepup of fifth harmonic voltage arising in other sources because of series resonance.

3. In a number of cases, large changes have been noted at irregular times in the severity of inductive exposures (usually involving rural power circuits) without any apparent cause. In several of them, these changes were finally associated with opening or closing of circuit breakers on the transmission line network, in some cases literally hundreds of miles away. Evidently, changes in propagation conditions on the transmission system resulting from the operation of the breakers were responsible.

4. In one of the first metallurgical rectifier installations to employ a large bank of capacitors, an attempt was made to determine the effects on inductive co-ordination of operation with certain rectifier units shut down in the presence of suspected resonance between the capacitors and the power system at one of the frequencies normally suppressed by phase multiplication. The attempt was abandoned because of difficulties in rectifier operation and danger of burning out the capacitors because of large currents at the resonant frequency.

5. One of the sharpest resonance cases encountered was between a rural line and the transformer feeding it. Here a modest 1,380-cycle voltage was stepped up to about 20 per cent of the rms; the 1,380-cycle current was larger than the 60-cycle current.

6. Very large fifth harmonic currents have been observed in power factor correction capacitors on distribution systems because of resonance, and some cases of capacitor overheating have been observed, mainly in installations by power company customers and because of resonance with the supply transformers.

7. One of the highest voltage TIF's on record, about 2,500, was caused by about a four to one stepup caused by resonance between a small generator and transformer and a rural power line.

Appendix III. Examples of Types of Situations Studied

In Table XII a few of the installations which have been studied are summarized. The examples listed there were chosen to represent a number of different types of situations.

Items *A* through *D* are typical of very large installations and are included to illustrate several factors, namely:

1. The large advantage of phase multiplication, for example, compare examples *A* and *A*[1], which are for the same installation before and after phase multiplication was applied.

2. The further reduction in influence obtained through the use of large banks of capacitors installed for power operating reasons (see example *B*).

3. Example *D* illustrates the type of situation which occasionally arises where some difficulty may be experienced even with 36 phases, particularly when one unit is out of service. In this situation, during the existence of the unfavorable power layout, the removal of units for maintenance was confined as far as practicable to periods of low telephone traffic.

Items *E*, *F*, *G*, and *H* are street railway rectifier installations chosen to illustrate a range of conditions. Item *E* illustrates a rather extreme condition in that while most factors involved were favorable, there was a condition of resonance between an extensive cable network and a large generating station which accentuated the seventeenth and nineteenth harmonics. Changing to 12-phase operation greatly reduced these harmonics and practically eliminated the problem. Items *F* and *G* are two similar street railway rectifiers installed in the same city and supplied over about the same length of the same type of feeder. The only difference was that in case *F* the feeder supplying the rectifier also supplied an extensive distribution system; in case *G* the feeder ended at the rectifier. The reason for the unusually large values of I·T in these cases is imperfect suppression of the seventeenth and nineteenth harmonics in the 12-phase arrangement. This condition was rather common in earlier rectifiers; it is seldom observed in more recent installations.

Item *H* is typical of many thousands of kilowatts of street railway rectifiers which have produced no problems on the a-c side but have occasionally produced problems on the d-c side.

Items *I* and *J* are radio station rectifiers and are included primarily to illustrate how the inductive effects are likely to be increased when the rectifiers are installed away from densely developed areas. A larger proportion of radio station rectifiers have caused co-ordination problems than any other class of installation because of this fact.

Items *K* and *L* illustrate again the difference in effects of relatively small rectifiers depending on location and power system conditions. Installation *K* is one of several in a very heavy industrial area and illustrates the type of situation where harmonic control would be desirable if cheaper methods were available; the influence and noise conditions are on the ragged edge, with little margin for future changes. Installation *L* is in a much less heavily developed area.

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Maintenance of Rectifiers on Electrochemical Installations

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Synopsis: Operating experiences and information gained over a period of years from the operation of a large group of mercury arc rectifiers have shown that, while rectifiers provide a very economical means of converting from a-c to d-c power, they are not entirely trouble free. The principal operating difficulty has been the seemingly inherent tendency for a rectifier to arc-back. This peculiarity has challenged both the designer and operator since it has been found that both contribute toward the solution. The present paper presents the features which have been found advisable for the operator to observe and control in order to keep the occurrences of arc-back at a minimum.

LARGE INSTALLATIONS of mercury arc rectifiers in the electrochemical industries of the United States made their first appearance in 1938 when they were installed to supply direct current necessary for the production of aluminum.

With the advent of the war and the demand for the light electrochemically produced metals, large amounts of direct current were required so that the development and installation of rectifiers was carried on very rapidly.

The details of the equipment[1] and the nature of the operating problems are well known.[2] Improvement in the operating characteristics of this relatively new apparatus is of interest to operating and design engineers. Possibilities for improvement, especially in the older or multi-anode design rectifiers, are disclosed in this paper covering the operating experience at one large installation. A broader view of remedies available possibly might be presented in discussions from other operating engineers, covering experience in irregular operation somewhat different from that of the installations on which this paper is based. Experience and observations reported cover equipment operated continuously at full load, which is peculiar to the electrochemical industry.

One of the oldest and most troublesome problems encountered in rectifier operation is the arc-back. Much research has been carried out on this problem and some improvement has resulted. In the 250-volt rectifiers, arc-backs are no longer a serious problem. With higher voltages the problem still exists to some extent, and there remains much to be learned about the cause and prevention of arc-backs.

It generally is agreed that the basic cause of an arc-back is the formation of a cathode spot on the anode. This spot destroys the valve action and constitutes a short circuit which must be cleared immediately. While there are several known causes for the formation of a cathode spot on an anode, a full explanation has not been found as yet, and it is believed that there are still many unknown causes.

In the various engineering publications, there is a wealth of highly technical data and discussion on arc-backs and their causes. Such data are available to all engineers and are not within the scope of this paper.

The equipment under consideration consists of rectifiers used in the production of aluminum. Two rectifiers are served by one transformer, which steps 13,200 volts down to 560 anode volts. The rectifiers covered are of three different types:

12 anode multianode type
12 tube ignitrons
12 tube excitrons

Figure 1 shows an inside view of one station of multianode rectifiers. Figure 2 shows one of these rectifiers with the anode plate assembly removed in order to repair a single anode.

The multianode rectifiers and the excitrons are of the grid controlled type. Voltage control is obtained by applying an adjustable negative potential to the grid, which correspondingly delays the firing of the anode. The ignitrons have a grid, but it is not used for voltage control. Ignitron voltage control is obtained by delaying the impulses to the ignitor. The main connections of a transformer and its rectifiers are shown in Figure 3.

Arc-backs are random in nature and are unpredictable. Some rectifiers have operated a year and a half without an arc-back, whereas others have had over a

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hundred in that length of time. There are not sufficient data available to say what constitutes too many arc-backs, or how many such mechanical shocks a transformer will stand. The shock the transformer receives is proportional to the value of the arc-back current, and the use of high speed anode circuit breakers reduces this current and, consequently, the transformer shock. As to frequency, the authors have used a sort of par or standard permissible rate of one arc-back per month per rectifier. This figure was based to some extent upon the rate existing at the time it was established, but subsequent experience has indicated that this figure can be reduced materially for single-anode per tank rectifiers.

A group of multianode rectifiers showed an over-all average of 1.8 arc-backs per month per rectifier during the first six years operation. The individual performance over this same period ranged from 0.6 to 3.0 arc-backs per month average.

From an operating standpoint, arc-backs may be divided empirically into two classes, and for the want of better terms may be called normal and abnormal arc-backs. Normal arc-backs are those which occur during the normal steady state operation and for which no cause is evident. Abnormal arc-backs are those which occur during or after some interruption to normal operation, such as removal from service or abnormal temperature or vacuum.

Some rectifiers, especially the multianode type, may have a good normal arc-back rate and a bad abnormal rate. Thus a rectifier may become sensitive or reach a condition wherein it operates better than par on steady service, but if taken out of service for a few hours it may arc-back a dozen times within five or six hours after having been restored to service. Conversely, some may have a poor normal rate but not be sensitive to such removal from service.

The cause for this behavior is not understood fully, but possible causes are the presence of condensed mercury in the vicinity of the anodes or the introduction of contamination from the anode shaft seals. The latter could result from the slight expansion and contraction of the shafts from temperature changes.

Operating Conditions

The majority of abnormal arc-backs are caused by contamination or leaks. Other causes are abnormal temperature or vacuum and excessively delayed firing for voltage control. Also, there are indications that some arc-backs are caused by droplets of liquid mercury or mercury dew on the anode or other surfaces at anode potential. Such arc-backs may occur after the rectifier has been out of service a short time and has cooled down somewhat. For example, there have been instances where the rectifier arced back as soon as the anodes were energized, even though there was no excitation or any d-c back feed. Sometimes four or five arc-backs would clear the trouble, but in other cases the bad anode would have to be disconnected for a few hours, after which it would operate normally.

Figure 1. Interior view of one station of multianode rectifiers

High temperature of cooling water tends to cause arc-backs but does not seem to be critical, because a good rectifier will permit a considerable increase of temperature without apparent bad effects. As an experiment a rectifier, whose normal jacket water temperature was 52 degrees centigrade, was operated about 24 hours at a temperature of 95 degrees centigrade without an arc-back. This rectifier, however, was one with a good record. A so-called sensitive rectifier very likely would arc-back under such treatment.

Shortly after a number of ignitrons were placed in service, a very prolific cause of arc-backs was found to be the splashing of mercury against the grid and anode. A small pocket in the bottom of the individual tank vacuum valves eventually filled with condensed mercury and overflowed down the pipe and against the baffle, and some of course went through the holes and against the anode itself. These arc-backs frequently set off one or more additional arc-backs. The accepted explanation of this phenomenon is that the concussion precipitates the overflow of the pockets in other valves. This trouble was corrected by installing internal baffles to prevent the mercury from reaching the anode.

Contamination

During the assembly of a rectifier, a certain amount of foreign material unavoidably is left in the tubes. Such material may consist of dust in the air, lint and dirt from workmen's clothing and wiping rags, residue from cleaning fluids, sand-blasting, and possibly other cleaning methods. New parts and the mercury itself may contain some impurities.

Some materials are much worse than others. Oil and grease, for example, are particularly bad, and their presence always will result in arc-backs. On the other hand, such materials as coarse sand, steel shavings, and broken porcelain insulation do not seem to cause much trouble. Even the lint from wiping rags and clothing does not cause as much trouble as would be expected.

In general, every effort should be made to keep the internal parts of rectifiers as clean as possible. The manufacturers and those industries that use a large number of rectifiers find it advisable to have a special air-conditioned room for opening and repairing rectifiers. Such a room should be controlled as to temperature and humidity.

Cleaning fluids, such as pure grade benzene (benzol), should be tested before using. An easy test is to evaporate about a pint and note the residue. If there is enough residue to be visible, it should not be used.

Where the various seals are made of materials such as rubber or asbestos, a certain amount of contamination may

come from these seals unless they are made suitably. During the war, for instance, some rubber gaskets were replaced by synthetic rubber. Under the influence of heat and vacuum, these gaskets liberated surprising amounts of compound into the tanks, and caused so many arc-backs that overhaul was required.

There has been no trouble which could be ascribed definitely to contaminated mercury. Such impurities as it may collect tend to rise to the surface where they are easily removed. A clean bright surface usually indicates satisfactory condition. It appears to be very difficult to make any laboratory tests which are any more reliable than visual inspection. Furthermore, the presence of certain kinds of foreign materials does not seem to cause any trouble.

As an experiment, several ignitrons were contaminated purposely with copper, tin, zinc, silver, and alumina (Al_2O_3), only one material, of course, being in each tank.

These materials have been in the rectifiers for periods ranging from seven months to a year and have a total of only two arc-backs for the five tanks. The only one examined to date is the one with the copper, which operated nearly a year without a single arc-back. Most of the copper was deposited conspicuously on the cooling coil and its support. The quartz ring in the mercury was definitely copper colored and a faint copper color was evident in spots on the grid. The scum on top of the mercury showed positive traces of copper, and there was some evidence of a slight amalgamation of the copper in the mercury, although this was not certain.

Mercury, which is removed from a rectifier cathode, always is given a standard cleaning treatment. This treatment consists of washing in nitric acid and distilled water, and then a double distillation is made. Mercury, which is suspected of containing oil, is given an extra washing with either grain alcohol or chemically pure benzol. The surface of clean mercury should be absolutely clear with no specks or film visible under a strong light.

Tests and Repairs

When the arc-back frequency shows a sudden increase or becomes what is considered excessive, the first step is to determine which anode or anodes are responsible. There are various devices and methods for indicating this action. One method, that has proved quite satisfactory, involves the use of a standard commercial surge-crest ammeter, which consists of a meter that measures the amount and direction of the magnetism in a special small magnetic link. One of these links is placed about six inches from each anode cable. The reverse current from the arc-back magnetizes the link correspondingly, so that the identity and magnitude of the arc-back can be determined by measuring the links after an arc-back with the surge-crest ammeter.

Figure 2. Assembly of cooling coil and anode plate for multianode rectifier

When such tests show only one anode in trouble, this one anode is usually repaired alone. If it is a multianode type rectifier it is opened and the one anode repaired. A special effort always is made to complete the work and close the rectifier up the same day, since with a short exposure time the degassing period can be shortened to a few hours.

Many of these single anode arc-backs are caused by leaks in the vicinity of the anode. This fact is especially true on seals consisting of organic material. Other causes are found in the anode assembly, such as broken grid insulation. Where there are two anodes in parallel, a broken grid wire on one will overload one anode, thereby tending to cause arc-backs. On the ignitrons a broken or wetted ignitor will have the same effect.

When the arc-back pattern is scattered over all or several anodes, and the rectifier is not a newly repaired one, an air or water leak is suspected first. Such a leak may be in the vacuum manifold, pumps, or some location an appreciable distance from an anode, so that its effect may be distributed over the whole rectifier. The easiest and first test for leaks is a hot pressure rise test, in which the low vacuum pump is stopped and the pressure rise noted for about an hour with the rectifier on normal load. A good rectifier will operate for hours with only a few microns increase in pressure. In fact, they have been known to operate a couple of weeks with the main valve closed and both pumps down. If there is a serious leak, the pressure should rise sharply in an hour or less. This pump stop test may be supplemented or confirmed by a cold pressure rise test. In this test the rectifier is cooled first to room temperature, and then the main vacuum valve is closed for several hours and the pressure rise noted. The temperature of the rectifier should be constant of course during the check period. The permissible rise for such a test varies with the types and sizes of rectifier tanks, but a rough figure is from three to six microns in six hours. Some tubes will show as little as one micron rise in this period, while others as much as 10 or 12 and still be acceptable. Some air leaks sufficient to cause arc-backs may be within the pump capacity and, consequently, not show up on the vacuum meter. These leaks will be indicated, however, by the foregoing tests, and in some cases can be measured with a low reading gas meter.

If these tests indicate that there are no

air leaks, then contamination, water leaks, or internal trouble is suspected, and the only course is to open the rectifier.

Air and water leaks leave a certain amount of discoloration at their source. Contamination usually leaves some discoloration or spots, and in some cases a definite deposit on surfaces near the source.

The inside appearance varies considerably with different rectifiers. On some the inside of the tank and the cooling coils have a dark smoked appearance, whereas others show various temper colors of blue and straw. Air leaks usually leave a light brown rust deposit on the metal surfaces. On others a great portion of the surface is white because of the amalgamation of mercury vapor.

The inside of the tank and the cooling coils are sandblasted. Finished steel surfaces are polished with emery cloth. All cleaned surfaces then are wiped with a clean lintfree rag and a cleaning fluid such as pure benzene. Insulating parts also are sandblasted and cleaned the same way.

Great care and cleanliness must be used in the assembly. A touch by the bare skin on a finished surface will produce a small rust spot which is a possible source of contamination. Clean cotton gloves and clothing must be worn while making the assembly.

In most cases the same cathode mercury possibly could be reused by skimming off the surface, but it is advisable to use only mercury which has been renovated thoroughly, as previously described.

In old rectifiers the insulating parts which are exposed always show a certain amount of metallization. This metallization is a gradual process in which metal from the steel parts is deposited on the various insulating parts. The actual causes and details of the transfer of metal does not seem to be understood thoroughly, but one theory is that the arc stream for some reason may be diverted from its normal path and travel through some steel part, such as the cooling coil or tank wall, thereby vaporizing some of the metal. Pitted spots have been found in rectifiers which indicates that this phenomenon occurs.

Under normal operation the metallization is quite slow and may require years to become critical. At the other extreme, there are cases where it became so bad within a few weeks that the rectifier had to be reopened and cleaned. In such cases there is always a reason for this condition which may or may not be discovered. Improper degassing operation, in which the pressure is allowed to run too high, excessive tank temperature, leaks, and contamination will hasten the process of metallization.

When this metallization has progressed far enough, the insulating value of the insulating parts is impaired to a point where improper operation results. Thus the grid and anode insulation may become so low that the grid will receive its potential from the anode instead of its regular source, and, consequently, lose its voltage control ability. In the final stages of metallization, the arc-back rate increases and the rectifier must be opened and cleaned.

Degassing

When a rectifier has been exposed to air, it must be degassed in order to remove the gas which was absorbed during the exposure.

The degassing time depends to a great extent upon the length of the exposure time. Thus a rectifier, which has been opened and closed the same day with an exposure of only a few hours, can be degassed in a few hours. On these short exposures, it is possible to degas in service on line voltage at low loads, and, while this practice sometimes is followed, it has been found better to give it at least a short period on short circuit degassing.

A "green" rectifier, that is, one which has been opened for several days, requires from 60 to 120 hours of degassing.

The common practice in degassing is to short circuit the rectifier by connecting the cathode to the transformer neutral (the negative line) and apply 40 to 80 volts to the anodes. On 13-kv transformers, for example, 1,700 to 2,300 volts on the primary makes a satisfactory degassing voltage for 600-volt class rectifiers. On grid control type rectifiers, there should be some resistance in series with each anode in order to insure proper load division. This resistance is of the order of one-half to one ohm at the start and is reduced as degassing progresses. At about three-quarters of the rated full load current, the resistance may be removed if desired and the degassing finished without resistors. The short circuit current is controlled by the regular voltage control method of either grid bias or ignitor phase shift, depending upon the type of rectifier.

Ignitrons may be degassed satisfactorily without anode resistors even when "green." Some users, however, insert a low resistance of about one-fourth ohm in each anode until the current has reached

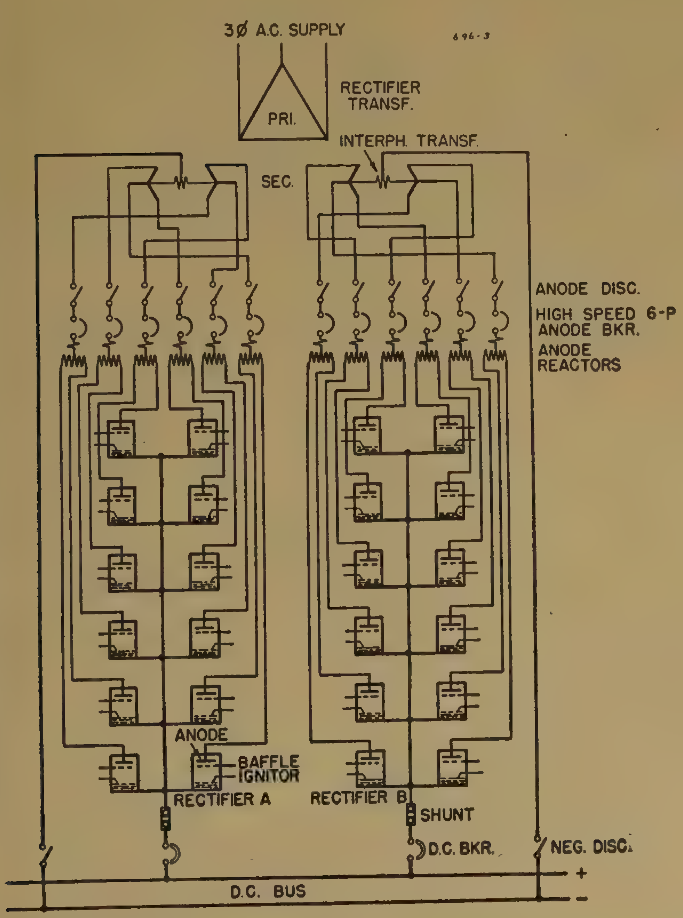


Figure 3. Unit connection diagram for ignitron rectifier

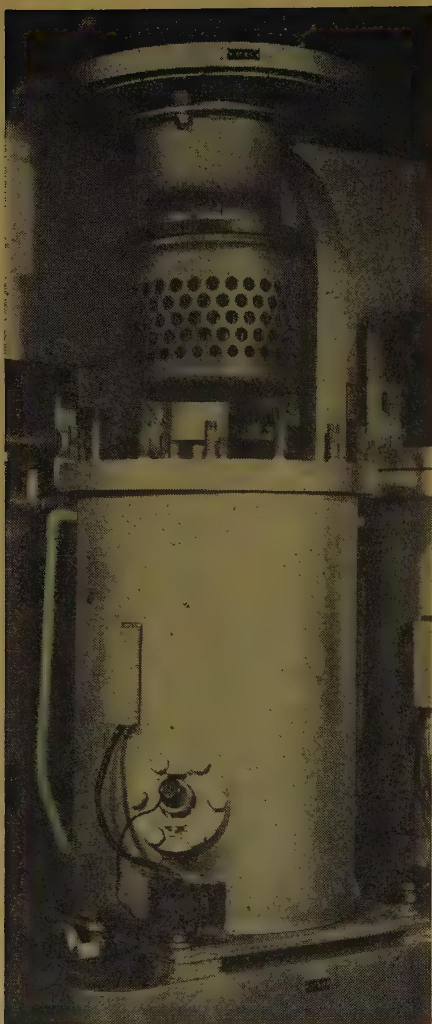


Figure 4. Anode and grid assembly of an ignitron

one-fourth to one-third rated value. This resistor balances the anode currents and results in a more uniform cathode current.

The degassing current is started at about two per cent of normal for multianode tubes and about ten per cent of normal for ignitron tubes, and increased as fast as the vacuum will permit. Before starting, the tube is pumped down to practically zero pressure on both the McLeod and electric gauges. During the degassing period, the vacuum on the multianode rectifiers is held within two microns on the McLeod gauge, and from four to ten microns on the electric gauge. The higher indication of the electric gauge during degassing is caused by the presence of condensable gases, which are not recognized by the McLeod gauge. After the rectifier has been degassed thoroughly, both gauges will settle down to approximately the same reading.

Ignitrons may be degassed with a somewhat higher pressure than the multianode type. McLeod readings of four or five microns and electric gauge readings of 12 to 15 microns are permissible on the ignitrons.

While the liberation of gas is fairly uniform during most of the degassing period, the gas occasionally will come out in sudden bursts or bubbles as though a pocket were released. When this action occurs, the degassing current must be reduced or removed to enable the vacuum pumps to restore proper vacuum. There are evidences that such pockets develop after the rectifier has been degassed and put back in service. For example, there have been several instances in which an arc-back on a recently overhauled rectifier was accompanied by sufficient increase of pressure to show on the vacuum gauge. This condition existed for only a few minutes.

After the regular degassing period, a rectifier occasionally continues to liberate a measurable amount of gas for several days, or even weeks after it is back in normal service. For electrochemical service it has been found advisable to finish a degassing operation by applying an overload current of 25 to 50 per cent for a few hours, in order to remove as much gas as possible. During the degassing period, a large rectifier will liberate from one-half to nearly one cubic foot of noncondensable gas. Figure 5 shows the time-gas volume curves obtained at Alcoa from the degassing of two "green" rectifiers selected at random.

In the degassing of "green" rectifiers, the gas is liberated most rapidly at relatively low current loads ranging from about 15 to 40 per cent of the normal operating load. Thus, in a typical degassing operation, the current is increased more or less uniformly until this critical value is reached. At the critical value, the gas is released as fast as the pumps can handle it, and the current may have to be maintained constant for several hours.

Having once passed this "hump," it progresses more rapidly as the load is increased. When a rectifier has been given a complete overhaul and placed back in service, it occasionally will arc-back several times during the first week or so, after which it settles down gradually to good performance.

It is believed that this behavior is caused by contamination left in the tube after overhaul. In order to hasten decontamination, a process called "aeration" has been tried several times at Alcoa. This process consists of cooling the rectifier and admitting clean dry air for some two or three hours. This air is admitted through a tube containing a filter and a drying agent, such as activated alumina. The air then is pumped out and the rectifier degassed on full voltage and a slowly increased load. The theory of this "aeration" is that a certain amount of the contaminants are oxidized and mixed with the air, and these products are pumped out sooner than they would be in normal operation. Engineers are not in full agreement on this theory, but favorable results have been obtained and the method should be given thorough trial.

Conclusions

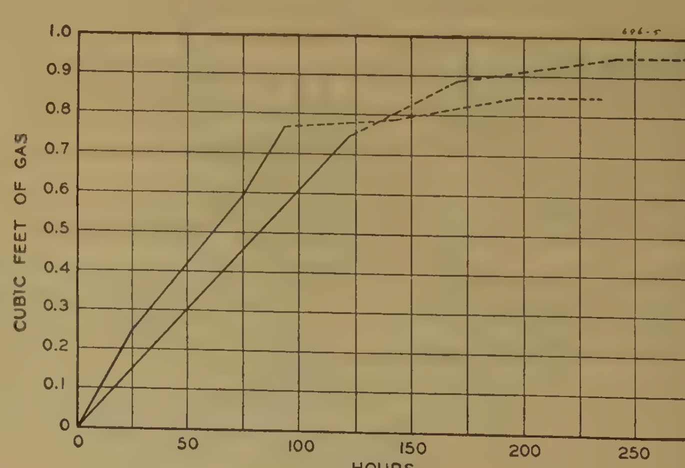
It generally is agreed that the single anode type, such as the ignitron and the excitron, is superior to the multianode type. Their efficiency is higher because of the lower arc drop, and their maintenance is simpler and cheaper.

As compared to rotating equipment, modern rectifiers with their associated auxiliaries require somewhat less maintenance although this maintenance does require more care and skill, especially in supervision.

Rectifiers are capable of a certain amount of reduction from full voltage by delaying the time in the cycle at which

Figure 5. Time-gas volume curves obtained from the degassing of two "green" rectifiers selected at random

——— Degassing
- - - - - - On line



the anode begins passing current. Such control is obtained at the expense of an increase in arc-back frequency, a lower power factor, and a slight decrease in efficiency. The greater the reduction, the more pronounced are these undesirable effects, so that reduction in excess of some 15 or 20 per cent is of doubtful value. For this reason it is preferable to operate with the minimum firing delay compatible with operating conditions. This voltage control feature stands in need of further improvement.

Inasmuch as a cold rectifier is more likely to arc-back, means must be provided to keep them warm. For continuous service this is no great problem, but for intermittent service suitable heaters usually are provided.

One of the most vulnerable points of a rectifier is the seals. These seals have been improved much in recent years, but there is still room for further improvement. This fact is especially true where they are made of organic materials, such as rubber and fiber.

There is not yet sufficient operating experience to say definitely how long a pumped type rectifier will operate before requiring a major overhaul, but the authors' experience would indicate a period of not less than six or seven years with continuous operation and uniform load. Aside from damage or failure of parts, the long time factor that should dictate a major overhaul is the metallization of parts or possibly an excessive amount of carbon in the mercury pool. Some minor repairs, such as ignitor renewal or seal repair, should be expected during the period.

From the user's standpoint, the outstanding problem is the cause and elimination of arc-backs. They are so unpredictable and so many occur without any known cause that they are a constant challenge to engineers and operators.

Among the known causes, contamination should have a great deal more study. There is much uncertainty as to the nature and amount of contamination necessary to cause serious trouble.

Since rectifiers are still on the threshold of their development, it is expected that future developments and research will simplify operation and reduce present maintenance problems.

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Fault Location and Relay Performance Analysis by Automatic Oscillographs

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Synopsis: The purpose of this paper is to describe a method developed for determining the location of ground faults on an extensive 140-kv grounded neutral transmission system. This method applies particularly to ground faults in a network of two or more parallel lines and is based on a comparison of the magnitude of ground currents in the faulted circuit and in the unfaulted circuits. Since the relative distribution of ground fault current in a network depends only on the zero sequence impedance of the system, an analysis of ground faults by comparison is independent of generation, fault resistance, or whether one or two phases are involved to ground. The paper also describes the use of ground fault current curves in determining the critical conditions for relay selectivity and summarizes the results of five years of satisfactory experience with automatic oscillographs used in the analysis of faults and relay operations on the system.

THE SYSTEM referred to in this discussion includes nearly 1,000 miles of line widely distributed over the State of Michigan, as shown geographically in Figure 1. The first automatic oscillograph was installed in 1931, and two more were added in 1935. These were used at various points on the system to study system behavior. In 1941 an additional one was acquired, and the four instruments were located permanently to co-ordinate fault records on the 140-kv grounded neutral system. In the period from 1941 to 1945 inclusive, oscillographic records were obtained for approximately 250 line faults. These records showed 96 per cent of the line faults involved either one or two conductors to ground. From this, it is obvious that fault studies on this system are concerned primarily with ground currents, and the location of the oscillographs and the element connections have been selected to give maximum information on ground faults.

Location and Initiation of Oscillographs

The instruments are located at the four principal neutral grounding points on the system: Croton Station on the Muskegon River, Morrow Station east of Kalamazoo, Blackstone Station at Jackson, and Weadock Station, Bay City. Initiation of the oscillographs is by instantaneous undervoltage relays connected to 140-kv potential devices and by instantaneous overcurrent relays in the 140-kv neutrals of the grounded transformer banks. The initiating relays are set to insure operation of at least two oscillographs for all faults within the network. It is a decided advantage in ground fault analysis to have films from two instruments for comparing current values or for determining the total fault current. In the event that one instrument fails to operate, a fairly satisfactory analysis may be made from one film.

Connection of Elements

The oscillographs in service are the 6-element type using sensitized paper. Except as noted, three elements of each instrument are used to record line-to-neutral voltage, and three are connected in the current transformer neutrals of the lines to record ground current. At the Morrow Station, four elements are required for residual current in the lines, leaving only two elements available for potential indication. These are connected to record line-to-line voltage.

Calibration

Records obtained on automatic oscillographs are no more accurate than the calibration data. In determining the values to be used for current calibration, it is first necessary to calculate the maximum value of fault current that may be obtained in the various elements, and to decide on the maximum amplitude that may be permitted on each element with maximum fault current. On elements centrally located on the paper, the total amplitude must be less than the width of the paper. On elements located near the top or bottom of the paper, it is not necessary to restrict the amplitude so that both the top and the bottom of the wave is confined to the paper under fault conditions. The zero line of the element can be drawn in readily and the total amplitude determined from the half wave which remains on the paper. For calibration tests a total amplitude of 12 millimeters on potential elements and 15 to 20 millimeters on current elements has proved satisfactory. A total amplitude of 12 millimeters is taken to mean six millimeters above and below the zero line.

Adequate account should be taken of the breakdown of current transformer ratios when measuring ground current, and in converting calibration test data into terms of primary ground current, the effective current transformer ratio must be used.

Valuable information has been obtained on relay and circuit breaker operating time by using a bias or offset on the current elements. This is obtained by connecting a shunt in the d-c trip circuit of the breaker to its respective element, causing the center line to shift when the trip circuit is energized. This is shown in Figure 7.

Factors Used in Ground Fault Analysis

1. Magnitude and relative values of ground currents from each end of the faulted circuit with both ends closed.

2. Magnitude of ground current in the faulted circuit after one end opens.

3. Relative values and direction of ground currents in unfaulted circuits in parallel with the faulted circuit.

Figure 1. Consumers Power Company 140-kv system (1945)

Paper 46-109, recommended by the AIEE committee on protective devices for presentation at the AIEE summer convention, Detroit, Mich., June 24–28, 1946. Manuscript submitted March 18, 1946; made available for printing May 1, 1946.

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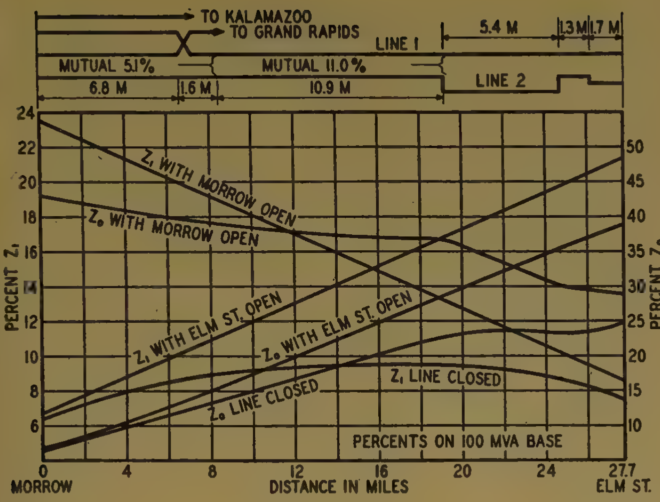


Figure 2. Curves of total fault impedance for line 1 and characteristics of line 1 and line 2 from Morrow Station to Elm Street Station, Battle Creek

(*a*). Both ends of the faulted circuits closed.

(*b*). After one end of the faulted circuit opens.

4. Magnitude and relative values of currents in contributing circuits from ground sources.

5. Recorded potential.

6. Changes in magnitude of ground fault current with changes in the nature of the fault, that is, one line to ground changing to two lines to ground.

7. Relay operation.

(*a*). Operating time of time delay relays.

(*b*). Pickup and operating range of instantaneous relays.

Calculations for Ground Fault Curves

Making up ground fault curves requires a considerable amount of work, but once completed, they provide a fast and accurate means of locating faults. A complete set of curves to utilize these factors consists of impedance curves, maximum ground fault current curves, and ratio curves.

The first step in making up data for curves is to have positive and zero sequence diagrams of the system and to calculate the positive, negative, and zero impedance for the total system at each station and at a sufficient number of points on the lines between stations to insure the drawing of accurate curves. On the average length of line, three or four points will be sufficient. Similar calculations are made for the lines open at either end. Ordinarily the positive and negative sequence impedances can be assumed to be equal. Calculations can be simplified by using linear values of impedance rather than the complex $R + jX$ form and are usually sufficiently accurate. If system generating conditions require the switching of grounded transformers which change the relative distribution of ground current, it may be necessary to make up more than one set of curves. As will be explained later, a second set of curves is not necessary for changes in generation only.

When calculating the total zero sequence impedance at each point, it is necessary to determine the distribution of ground current, not only in the faulted circuit but also in the parallel circuits and other contributing circuits. This is especially important for parallel circuits having mutual relation with the faulted circuit. For systems not involving too many mutuals, these values may be calculated by setting up mutual equivalents[1] or by simultaneous equations. It is important that fault points be taken at the ends of mutual sections, as the ground current curves change their slope at these points. On complicated systems with multicircuit mutuals, it is advisable to use an a-c calculating board.

Values of the total positive sequence impedance are more easily calculated or can be obtained on a d-c board.

Application of Curves

The two 140-kv circuits between Morrow Station and Elm Street Station, Battle Creek, designated as line 1 and line 2, have characteristics as shown at the top of Figure 2. A complete set of curves for line 1 is reproduced for illustration.

Impedance Curves

Curves shown on Figure 2 are for maximum positive and zero sequence impedances with the line closed and with the line open at either end. These data are convenient for checking a fault location when the nature of the fault changes. For example, assume a single-line-to-ground fault changes to a two-line-to-ground fault. If positive and negative

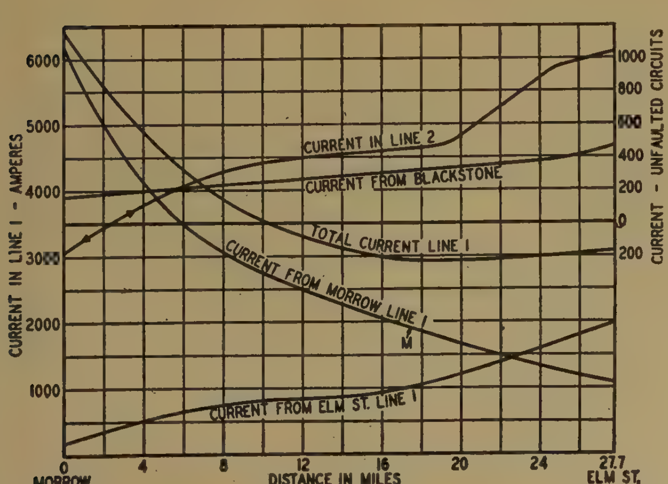


Figure 3. Maximum current curves for single-line-to-ground faults on line 1, closed at Morrow and Elm Street

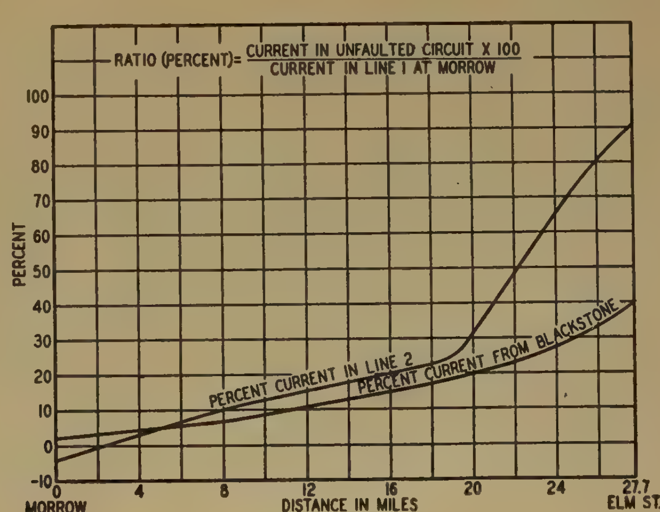


Figure 4. Ground current ratio curves for faults on line 1, closed at Morrow and Elm Street

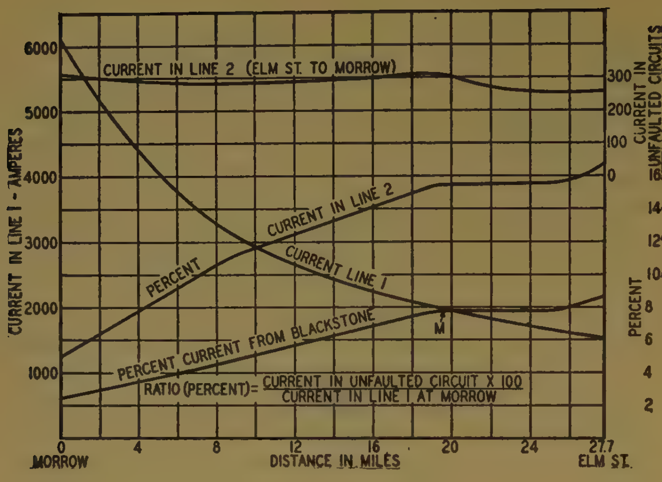


Figure 5. Maximum ground current curves and ratio curves for faults on line 1, open on Elm Street end

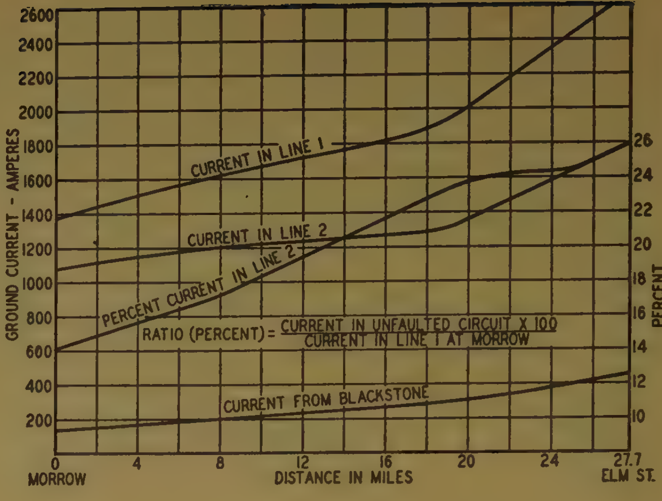


Figure 6. Maximum ground current curves and ratio curves for faults on line 1, open on the Morrow end

phase impedances are considered equal,

Single-line-to-ground current

$$=\frac{3\times 10^7}{(2Z_1+Z_0)\sqrt{3}\text{ kv}} \quad (1)$$

Two-line-to-ground current

$$=\frac{3\times 10^7}{(Z_1+2Z_0)\sqrt{3}\text{ kv}} \quad (2)$$

A value of Z_1 and Z_0 for the same location can be taken from the impedance curves which when applied in the above formulas will be equal to or proportional to the current values shown on the oscillograph film. This is usually a good indication of fault location, especially on a line section where there is a large variation in the ratio of Z_1/Z_0.

Current Curves

Current curves may be used directly in determining fault location if the fault currents are near maximum values. The curves are also very useful in determining the critical conditions for relay selectivity and for co-ordinating the magnitude and duration of fault current with observed line damage.

Maximum current curves for single-line-to-ground faults with the line closed at both ends are shown in Figure 3.

Curves for line 1 open at Elm Street and open at Morrow are shown combined with ratio curves in Figures 5 and 6, respectively. Note in Figure 3 that the current in line 2 passes through the zero line at about two miles from Morrow Station. That is, for faults within two miles of Morrow, the current in line 2 is from Elm Street to Morrow, and for faults beyond two miles, the current in line 2 flows from Morrow to Elm Street. Whether or not the current in a parallel circuit reverses after a circuit breaker opens on one end of the faulted line is often a definite indication of the fault location.

The difference in the slopes of the current curves in the 10.9-mile section and in the 5.4-mile section clearly indicates that the effect of mutual impedance cannot be neglected.

Ratio Curves

The distribution of ground fault current depends only on the zero sequence impedance of the system. The relative distribution in the faulted circuit and in parallel or contributing circuits is the same for one line-to-ground or for two line-to-ground faults. Fault resistance or changes in generation affect the magnitude of ground fault current but do not change the relative distribution. This fact can be utilized in making up a set of ratio curves independent of fault resistance and generation. Using the current in one circuit as a reference, the current in the other circuits can be plotted in per cent of the reference current.

It is apparent that fault locations can be determined more accurately if the ratio curves have a steep slope. On networks having only two currents to compare, it is not always possible to have one ratio curve that will have adequate slope over the entire length of line. However, nearly all line faults are cleared by the sequential opening of breakers at each end. This supplies data for making up three ratio curves, one of which can be used for accurate indication of fault location on a specific section of the line.

On more complicated networks where three or more currents can be compared, it may be advisable to make up two sets of ratio curves for the same condition by using different currents as reference. That is, by proper selection of reference currents, one curve may have a steep slope for a section of the line where the other curve is relatively flat.

Ratio curves apply specifically to faults with reduced values of current, but they can be applied equally well to faults with maximum current values. In some cases the ratio of currents may give a more definite indication than the individual currents.

The ratio curves of Figure 4 are obtained by plotting maximum current values, taken from Figure 3, in per cent of the current in line 1 at Morrow. Ratio curves with the line open at Elm Street and at Morrow are included in Figures 5 and 6.

Application of Oscillograph Data for Fault Location

The oscillograph films, Figure 7 and Figure 8, were obtained from Morrow Station and Blackstone Station respectively for a fault on line 1.

A preliminary inspection of the films shows the two line-to-line potentials at Morrow are approximately equal, indicating that the Z phase was involved. Reference to the Blackstone line-to-neutral potentials shows definitely that the fault was Z-phase-to-ground throughout. Further inspection of the film shows that the line opened at Elm Street in 19 cycles and at Morrow in 48 cycles. Targets were reported on the time delay ground relays at both Elm Street and Morrow. As indicated at M in Figure 5, the maximum reach of the instantaneous relays on line 1 at Morrow after the line opens at Elm Street is approximately 19

miles. Failure of this relay to operate indicates that the fault was more than 19 miles from Morrow or within nine miles of Elm Street. The bias was applied when the trip circuit at Morrow was energized. This shows that the oil circuit breaker on line 1 had five cycles clearing time.

Use of Current Curves

A detailed analysis gives measured values of fault currents in primary amperes as shown in Figures 7 and 8. The current values of the fifth and sixth elements of both films are not required in this analysis.

No film was obtained from Weadock for this fault, but previous calculations determined that for a fault at Elm Street, 51 per cent of the ground current from the eastern part of the system is contributed over the line from Blackstone, and 49 per cent over the line from Lansing. From the measured and calculated values of current, the distribution in the lines was as shown in Figure 9A for the initial fault and in Figure 9B after Elm Street opened. This gives a complete picture of the fault and indicates that the calculated total of 2,965 amperes was very nearly maximum current for a fault near Elm Street.

Applying the values shown in Figure 9A to the maximum current curves of Figure 3, the 1,400 amperes in line 1 from Morrow gives a fault location of 4.5 miles from Elm Street; 860 amperes in line 2 gives a location of four miles from Elm Street; 360 amperes from Blackstone gives a location of 5.5 miles from Elm Street; and the calculated value of 1,565 amperes in line 1 from Elm Street gives a location of 4.5 miles. This close agreement on fault location is a further indication that the fault currents were very nearly maximum value.

After the line opened at Elm Street, the values shown in Figure 9B applied to the current curves of Figure 5 give a location of five miles from Elm Street for 1,750 amperes in line 1, and 5.5 miles from Elm Street for 270 amperes in line 2. Note that the direction of current in line 2 reversed after line 1 opened at Elm Street. The curve of total amperes on Figure 3 is too flat at this point for close application, but the calculated total indicates the fault as being between four and six miles from Elm Street. All of the above determinations give the fault location as between 4 miles and 5.5 miles from Elm Street. The actual location was five miles from Elm Street, and the damage reported was three broken bells and light burns on the conductor.

Use of Ratio Curves

To illustrate the application of the ratio curves, assume that fault resistance or decreased generation had reduced the fault current to 80 per cent of the above values. Relative values then would be as shown in Figures 10A and 10B. Applying the values in Figure 10A to the current curves of Figure 3, the 1,120 amperes in line 1 from Morrow indicates the fault to be at or near Elm Street, while 690 amperes in line 2 places the fault at six miles from Elm Street, and 290 amperes from Blackstone gives the location as 14 miles from Elm Street. Similarly, the values of Figure 10B applied to the current curves of Figure 5 all indicate the fault as being at or near Elm Street. This wide discrepancy in the indicated fault location is characteristic of current values that are less than maximum and is an indication that the current curves cannot be directly applied, and the ratio curves should be used.

From the values of Figure 10A, the current in line 2 from Morrow is 61.5 per cent of the current in line 1 from Morrow, and the current from Blackstone is 26 per cent of the current in line 1. Applying these percentages to the ratio curves of Figure 4 gives fault locations of 4.5 miles and 4 miles from Elm Street as previously determined with actual fault currents. The currents in Figure 10B, after Elm Street opens, applied in the same way gives 15.5 per cent for the current in line 2 and 7.8 per cent for the current in the Blackstone line. These values fall on the flat part of the ratio curves of Figure 5 and serve only as a check.

From an inspection of the ratio curves of Figures 4, 5, and 6, it is evident that for faults within nine miles of Elm Street, the most accurate determinations can be made from the curves of Figure 4. For faults within 19 miles of Morrow, any one of the three sets of curves could be used with sufficient accuracy.

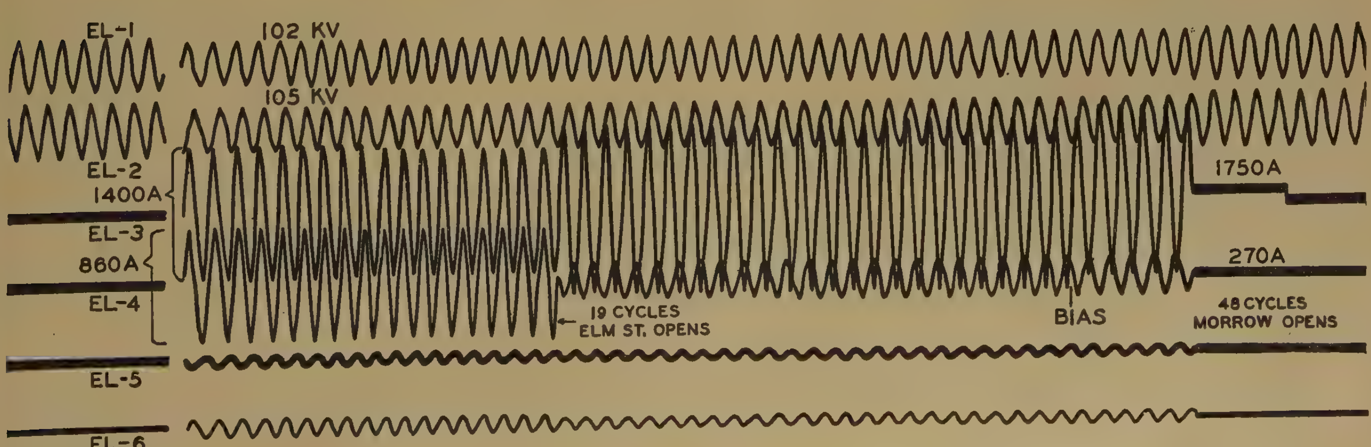
Other Applications of Oscillograph and Curve Data

Line Damage

During the past two years each fault analysis included calculations to correlate fault current and clearing time with the observed damage. In these calculations a factor has been used equal to (amperes/1,000)^2 $\times$ cycles. The currents used are taken from the oscillograph films supplemented by calculations and values from the current curves where necessary. For the actual fault discussed in this paper, the values used were 2,965 amperes for 19 cycles and 1,750 amperes for 29 cycles: $(KA)^2$ cycles$=255$. While this study has not been carried on for a sufficient length of time to warrant any definite conclusions, it is indicated that the data may be

Figure 7. Oscillogram from Morrow Station for a fault on line 1, Morrow to Battle Creek

Element 1. *YZ* phase potential
Element 2. *ZX* phase potential
Element 3. Ground current in line 1 to Battle Creek
Element 4. Ground current in line 2 to Battle Creek
Element 5. Ground current in Kalamazoo line
Element 6. Ground current in Grand Rapids line



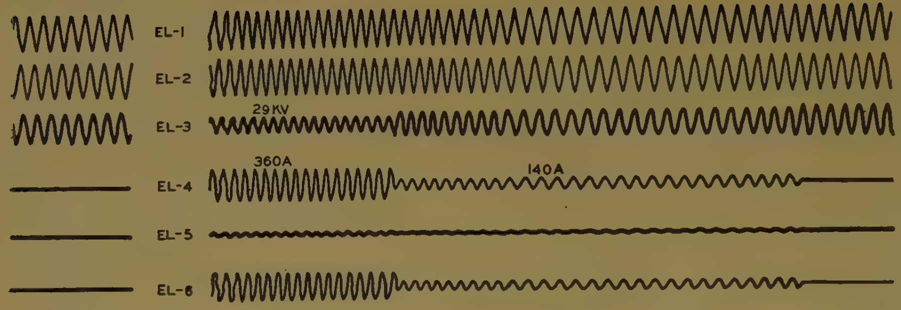


Figure 8. Oscillogram from Blackstone Station, Jackson, for a fault on line 1, Morrow to Battle Creek

Element 1. X-phase-to-neutral potential
Element 2. Y-phase-to-neutral potential
Element 3. Z-phase-to-neutral potential
Element 4. Ground current in line to Battle Creek
Element 5. Ground current in line to Lansing
Element 6. Ground current in interconnection

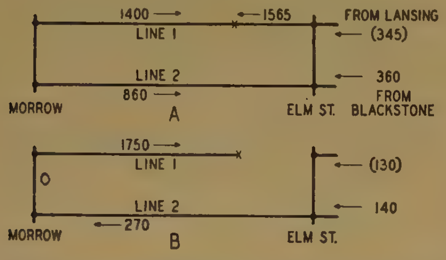


Figure 9. Distribution of maximum ground fault current

A. With line 1 closed at Morrow and Elm Street
B. After line 1 opened at Elm Street

of considerable value in determining whether or not a special patrol is necessary when no damage is found on the initial patrol.

Relay Performance

As a final step in the analysis, the operating time of each time delay ground relay is calculated for the value of ground current as measured on the films, and this calculated time is compared with the actual operating time as recorded. This furnishes a positive check on relay operation. In case of faulty films, this process can be reversed to calculate fault currents. Surprisingly accurate determinations have been made based entirely on relay operating time. Occasional faults

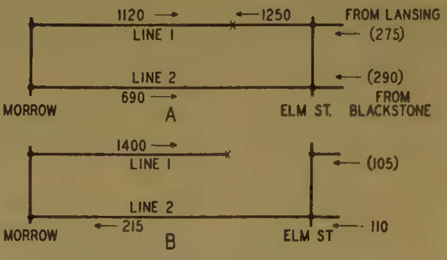


Figure 10. Distribution of ground fault current, 80 per cent of maximum

A. With line 1 closed at Morrow and Elm Street
B. After line 1 opened at Elm Street

near the end of the instantaneous relay zone supply data on the accuracy of pickup adjustment.

In the preparation of ground relay settings, the current curves are very useful in determining the critical points of selectivity. In many cases the critical point is at the end of the instantaneous relay zone where the time delay relay must be considered. This is especially true on parallel circuits with mutual couplings.

It can be seen in Figure 5 that the effect of the mutual is to maintain an almost constant value of current in line 2 for faults in the parallel section. And it is evident that faults on line 1 over 20 miles from Morrow may be critical for selectivity between the Elm Street relay on line 2 and the Morrow relay on line 1. Selectivity can be checked quickly and accurately by taking current values from the curves at any questionable point.

Summary

When making analyses of faults, it is of interest and value to prepare sheets for recording data on each fault. Tables I and II summarize the types and cause of

Table I. Types of Line Faults

Faults starting 1-line-to-ground	78%
Faults starting 2-line-to-ground	18%
Faults starting phase-to-phase	4%
Faults starting 1-line-to-ground and changing to 2-line-to-ground or to 3-phase	10%

Table II. Cause of Line Faults

Lightning	80.0%
Interference (birds, machinery, etc.)	9.0%
Wind	7.4%
Line material failure	1.2%
Unknown	2.4%

256 line faults analyzed over a period of five years. Exactly 50 per cent of these faults caused no visible damage and could not be located.

Conclusions

This method of determining ground fault location as developed by one of the authors has been in use for several years and has proved very satisfactory. For this discussion the particular line section and fault were selected to illustrate best some of the principles involved, and it should not be presumed that the same degree of accuracy can be obtained for all faults. With the present number of oscillographs on the system, the average degree of accuracy is in the order of ten per cent for faults in the network; that is, on a line 30 miles long it is expected that the faults can be located within three miles. With oscillographs at each station, it is probable that faults could be consistently located within one mile. The calculated locations of ground faults on radial lines are accurate only if tower footing and fault resistances are low.

One of the most important applications of the oscillograph is the information that is obtained when a fault is not cleared properly. In most cases these data definitely point to the cause of incorrect operation or failure to operate so that immediate steps can be taken to prevent a recurrence. By indicating the location of inconspicuous damage and potential sources of trouble, the oscillographs are becoming an increasingly important factor in transmission line maintenance. The diligent use of oscillograph data has resulted in greatly improved relay performance.

Reference

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Modern Diesel-Electric Drilling Rig Equipment

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Synopsis: A modern Diesel-electric powered drilling equipment for drilling oil wells consists of several internal combustion engines driving d-c generators, which supply adjustable voltage power to the several d-c motors driving the various motions. This paper gives a description of the functions performed by the draw works, mud pumps, rotary table, and coring reel, and a typical arrangement of apparatus or elementary power diagram showing the main d-c circuits with the arrangement of drill-hoist contactors and manual transfer switches. In addition, there are related the general scheme of operation, strictly from the driller's point of view, without the details of how these functions are accomplished; the operating principles of the electrohydraulic governor that make this scheme of operation possible; as well as the various control features provided by the amplidyne excitation of the main generators. The electrohydraulic governor, a war development, is applied for the first time to Diesel-electric drilling rig equipment. This governor gives to the engines a degree of protection never before attained, by preventing the engine from being overloaded. The governor also permits running the engines at the lowest speeds consistent with the desired rig motion and not at full rated speed at all times as in the past. The electrohydraulic governor provides for engine paralleling without the driller having to pay any attention to the engine throttles. The merits of a-c versus d-c auxiliary drives are pointed out, and separate engine-driven auxiliary a-c generators are recommended instead of the usual practice of driving auxiliary d-c generators from the main engines. This gives a saving in first cost of auxiliaries conservatively estimated at 25 per cent, as well as gains in simplicity, sturdiness, and less weight.

A MODERN Diesel-electric drilling rig equipment for drilling oil wells consists of several internal combustion engines driving d-c generators. These generators supply adjustable voltage power to the several d-c motors driving the draw works, mud pumps, rotary table, and coring reel.

Most oil well drilling is done by steam engines or Diesels that mechanically drive the various motions through a system of gears and clutches. However, oil wells are being drilled deeper in search of new sands, requiring more power. As in surface transportation and propulsion, the flexibility of transmitting this power electrically is becoming more and more evident.

It is estimated that approximately 35,000 wells will be drilled in 1946. The average depth will be about 4,500 feet. However, a great deal of the exploratory drilling between 10,000 and 15,000 feet now is being done with Diesel-electric equipment where its flexibility is of particular advantage.

It is the purpose of this paper to describe a typical modern Diesel-electric powered drilling equipment and not to discuss the relative economics of Diesel-electric versus mechanical and steam rigs.

Description of Functions

A modern deep well is drilled by means of a bit attached to a drill stem (drill stem is special hollow pipe) and rotated to bite through the sand, gravel, and rock formations of the earth's crust. The diameter of the hole may be 12 inches at the top and six or eight inches near the bottom of a 10,000- or 12,000-foot hole. For clarity, the functions of the various drives are given.

Draw Works

This equipment is the hoist for lifting the drill stem in and out of the hole, and a good view of it is shown in Figure 1. Bits have to be changed every day or two and sometimes every few hours. The draw works operates like a hoist with several mechanical gear changes. To change a bit, a section or two of drill stem is lifted out of the hole, unscrewed, stacked in the side of the derrick, and the traveling block and hook is lowered to pull up the string of drill stem another couple of sections, and so on. Usually, all engines are required for this operation while the mud pumps, rotary table, and coring reel remain idle.

Rotary Table

The rotary table rotates the drill stem with the bit on bottom and drills the hole. The driller regulates the weight of pipe permitted on the bit by means of a mechanical brake on the draw works. The chips and sand are removed by circulating mud.

Mud Pumps

The principal functions of the mud pumps are to circulate the mud which lubricates the drill stem and bit, pick up the chips and sand, and bring them to the surface, suspended in the mud, in the annular space between the drill stem and the outside steel casing, as illustrated in Figure 2. Getting the mud to the right consistency to do this job is a "black" art known only to the drillers. Keeping the mud flowing is important because a 10,000-foot column of mud has a burden of chips which will settle out at the bottom and may freeze the drill stem in the hole if left to stand for a few hours.

Coring Reel

This reel is a light-duty high-speed hoist used to remove sample cores from the hole and for logging purposes. It is used when other apparatus is not in use and is, mechanically, entirely independent.

Typical Arrangement of Apparatus

A typical arrangement of apparatus for 15,000 foot drilling is shown in Figure 3. In this example, three Diesel-engine-driven generators are shown. In some cases, four, five, or even six such sets are employed. However, the minimum rating of Diesel-engine set often is dictated by the rating of the mud pump so that one engine generator set can supply its full rating. The number of such sets is dictated by the rating of the draw works.

The auxiliary motor-driven apparatus, such as miscellaneous pumps and blowers, and the power-system used to supply it will be discussed later.

General Scheme of Operation

Perhaps it will give a better understanding of the equipment operation to describe the results obtained by modern electric equipment from the driller's point of view, before describing in detail the equipment used for accomplishing the results.

Each of the motors operates at constant field strength. Speed is controlled entirely by generator voltage (that is, by generator field and engine speed).

Assume that on Monday morning all equipment is at rest. The engines are started and idled at half speed. From this part on, all control of the rig by the driller

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Figure 1. General view showing draw works, swivel with mud-hose connections, hexagonal kelly, and rotary table

is from the four driller's speed master switches and the selector switch shown in Figure 3. To illustrate how the driller's control works, observe what the driller does to start a mud pump. An engine-generator set previously has been assigned, by means of the manual transfer switches, to supply power to this particular mud pump.

The driller moves the speed master switch handwheel from the off position which starts the mud pump. The further he moves the handwheel, the faster the mud pump runs. Up to approximately half motor speed, the engine remains at its half speed. As the driller calls for more motor speed by moving his speed master switch further, the engine automatically speeds. The driller does not have to touch the engine throttle, nor worry about overloading the engine. If the mud pump should become overloaded or even stalled, the load on the engine will be reduced so that the engine is not harmed. These functions are accomplished automatically by means of the engine governor and amplidyne control, which will be explained later.

If desired, the control equipment can be preset to limit the stalling torque on the motor from 75 to 150 per cent of normal motor torque. The engine may be operating at full speed and the mud pump suddenly stall, caused by a plugged bit at, let us say, 125 per cent of normal torque, regardless of where the driller's master switch may be. This torque limit adjustment is made at the control cabinet and not at the drillers position.

To stop the mud pump, the driller returns the master switch to the off position. The Diesel engine returns to idling, at half speed, and the mud pump motor stops.

The controls for all the other motors are the same. Reversing a particular motor merely requires the master switch to be returned to the off position, the handwheel pulled out and rotated as before. The motor reverses and accelerates in the opposite direction. Reversing is not required for mud pumps but is available.

Consider a few of the typical driller's operations.

"Spudding-in"

"Spudding-in" is a process of starting a hole and drilling it the first several hundred feet. There is not sufficient weight on the bit to make it bite at this stage. It is necessary for the driller to lift the drill stem a few feet and then drop it, thus giving the bit a fresh bite. Then it is rotated and may be done at the rate of several times per minute.

The driller has a common speed master switch for the draw works and the rotary table, as shown in Figure 3. It will operate either, but not both simultaneously. There is a "drill-hoist" selector switch which connects generators alternately to the rotary table or to the draw works, provided the speed control switch is in the off position.

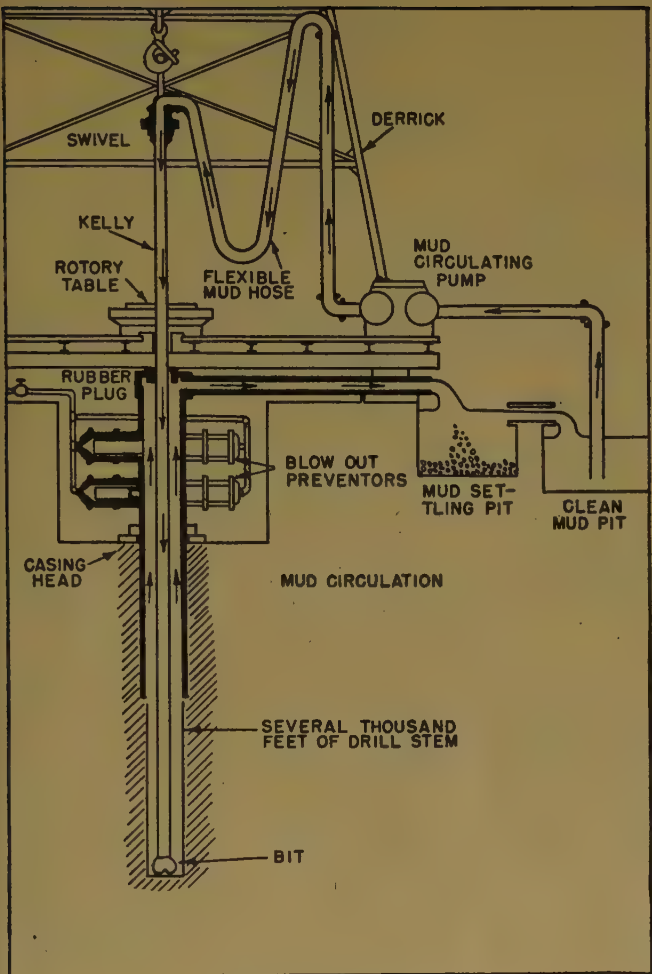
The driller turns the drill-hoist selector switch to the *DW* position and turns the speed control master switch, raising the drill stem a little way. He then returns the master switch to the off position and declutches the motor from the hoist drum (the clutch is a pneumatically operated device) and lets the tool drop.

The drill-hoist selector switch is turned to the opposite position to *RT*, and the speed master is turned to rotate the table for a few revolutions. Then the process is repeated.

During this operation, one mud pump motor and engine set has been operating, probably at low speed, circulating mud. This engine set has been prevented from transferring in response to the drill-hoist selector switch by means of a selector switch on the control board.

Drilling

The "drill-hoist" selector switch is thrown to the drill position, which con-



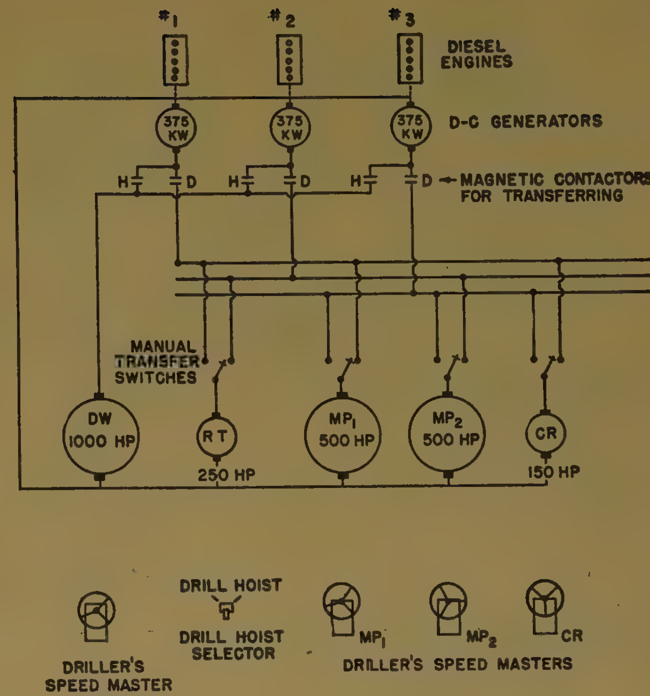


Figure 2 (left). Cross section of rig showing how mud pump circulates mud

Figure 3 (above). Elementary power diagram of typical d-c drilling rig equipment

nects one engine set to the rotary table and another engine set to a mud pump. The third engine may be used on the second mud pump or it may be shut down.

The driller starts and runs the rotary table, at the desired speed with the speed master, and has no throttle control for the engines, since this is under the control of the engine governor.

Hoisting

When it becomes time to change the bit on the end of the drill stem, the mud pumps are shut down first and the flexible mud hose, swivel, and kelly assembly disconnected from the drill stem. Meanwhile, the engines are idling at half speed because all of the speed masters are in the off position.

The mud hose swivel and kelly (the square piece of drill stem that slides up and down in the square hole in the rotary table) then is dropped in the "rat hole." The "rat hole" is a spare hole that the driller thoughtfully has dug in advance for storing the kelly. It can be discerned in Figure 1.

The drill-hoist selector switch is thrown to the hoist position, which connects all of the engine-generator sets to the draw-works motor for maximum power. All of the engines are paralleled automatically and each may be loaded in accordance with its ability to carry load. The driller turns the speed master in accordance with the desired hoisting speed. If the string of drill stem is exceptionally heavy, the motor will slow down automatically to keep its input within the engine's ability, regardless of the speed-master position. If the drill stem should get stuck, the draw-works motor would stall at about 200 per cent torque or some less adjustable value without damage to the engines, generators, draw-works motor, or to the mechanical equipment, even if the speed master is in the maximum speed position. However, the driller should not let this condition persist for any period of time, since the thermal capacity of the draw-works motor does not permit it to develop 200 per cent torque continuously.

After the drill stem has been hoisted three lengths (a "thribble"), the driller returns his speed master to the off position and sets his hand brake. The drill stem remaining in the hole is wedged carefully into the rotary table. The driller then unscrews the withdrawn lengths of drill stem by holding the withdrawn portion with tongs and rotating the rotary table by power.

After the withdrawn drill stem section is stacked over in the derrick, the hook is driven down at high speed and light load to get another grip on the drill stem remaining in the hole. For this operation, the driller's speed master is pulled outward and turned as far as it will go to get maximum speed.

It may require several hours to take the drill stem out of the hole, attach a new bit, and lower it into the hole again, section by section. This process is called a "round trip."

In lowering a heavy load, however, the draw-works motor is declutched and the lowering energy is dissipated in a water-cooled brake. The Diesel engine cannot absorb very much energy and regenerative lowering has not been resorted to as yet.

In all this maneuvering, the driller needs pay no attention whatever to anything but his speed master and drill-hoist selector switch. All paralleling, speeding up, and slowing down of engines is entirely automatic. When they are paralleled, as on hoisting duty, there is no delicate adjusting of throttles to contend with or nursing of a smaller or defective engine. The automatic governors will make the engines take all the load of which they are capable, and no more, and make each engine parallel with the others.

Manually Selected Combinations

It may be necessary to shut down an engine-generator set for maintenance purposes or simply because its power is not needed. It is desirable, therefore, to give each main drive motor two choices of

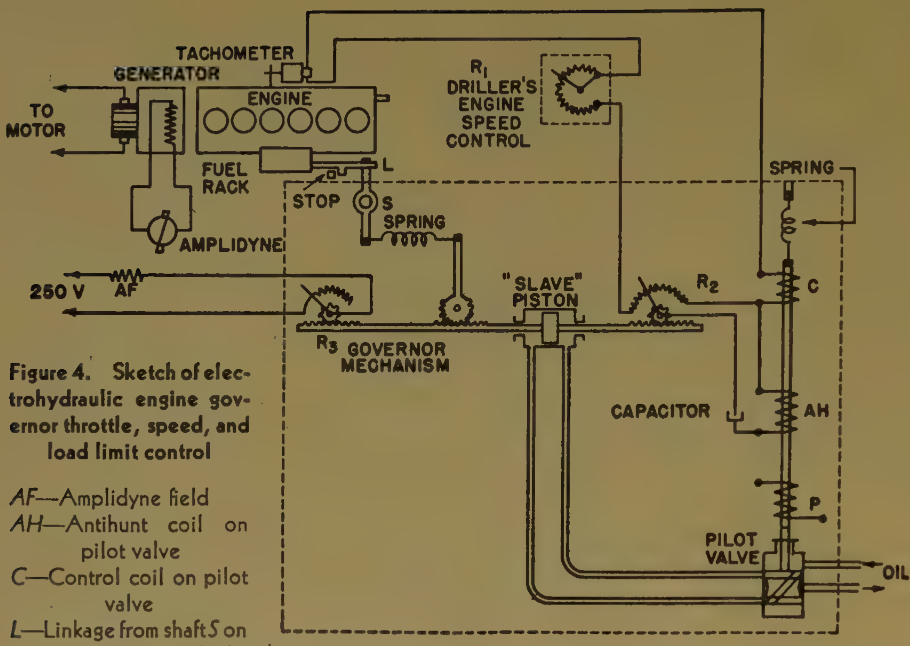


Figure 4. Sketch of electrohydraulic engine governor throttle, speed, and load limit control

AF—Amplidyne field
AH—Antihunt coil on pilot valve
C—Control coil on pilot valve
L—Linkage from shaft *S* on governor to engine fuel rack
P—Coil on pilot valve for paralleling engines
R_1—Driller's speed control rheostat
R_2—Antihunt rheostat
R_3—Rheostat load limit
S—Governor shaft for connection to fuel rack

engine-generator sets. This choice is provided by the manual transfer switches on the control panel.

A selector switch for each engine, located on the control panel, will take away control from the driller's drill-hoist selector switch so that an engine may be retained on some special duty, such as driving a mud pump for mud mixing even while hoisting.

Electrohydraulic Engine Governor

The internal combustion engine is essentially a constant-torque device over an adjustable speed range. It rapidly overheats when submitted to overloads with resultant increase in maintenance.

Since a drilling load is both intermittent and varying, it is highly desirable to run the engines only at the lowest speed consistent with the power required. This rate of speed saves both fuel and wear and tear on the engine. To appreciate this point, just imagine the maintenance and fuel consumption on an automobile engine if it is run always at top engine speed, and the car started and stopped at each traffic light with the clutch and gear shift. That procedure had been followed by the drilling industry with Diesel-electric drives up to the advent of the electrohydraulic governor.

The fuel injection system on most Diesels is such that at low speeds, full throttle opening will overheat the engines. That is, with the throttle wide open and the engine at low speed, actually more fuel per stroke is injected than at full rated engine speed. Thus a requirement of the governing system is to adjust automatically the throttle stop as a function of engine speed.

Functions of the Governor

The functions of the governor can be summarized as follows:

1. It measures engine speed and adjusts the engine throttle to maintain this speed up to permissible full throttle opening.

2. When the throttle is fully open and a further attempt is made to increase load on the engine, it automatically reduces the generator field.

3. It automatically adjusts the position of the throttle stop as a function of engine speed.

All these requirements are accomplished rather simply by the electrohydraulic governor. Refer to Figure 4.

An electric tachometer produces a d-c signal proportional to engine speed. This signal is put into a solenoid operated pilot valve. The pull of the solenoid *C* is balanced against a spring. A slight unbalance in engine speed thus permits the pilot valve to let in a little oil, under pressure, to one side or the other of a "slave" piston, which can move as far as necessary in adjusting the throttle to restore the engine speed and thus the balance between solenoid *C* and the spring.

These factors are the elements of any governing system. Since the slave piston has no restoring spring and can take up any position necessary to balance the solenoid *C* against the spring, the system is isochronous. This condition means that the system has no engine speed droop whatever as the load is applied.

The engine speed controller R_1 is a rheostat inserted between the tachometer and the solenoid *C*. The more resistance is cut in, the faster the engine has to run to establish the balance between solenoid *C* and the spring. The combination of pilot valve, "slave" piston, and throttle will see to that condition. Thus the engine will run at adjustable constant speed levels, depending on the position of rheostat R_1. It is usual to run the engines over either a 2/1 or a 3/1 ratio of speed range. The engine speed set by rheostat R_1 is entirely independent of the load the generator tries to put on the engine. R_1 is a part of the drillers speed master.

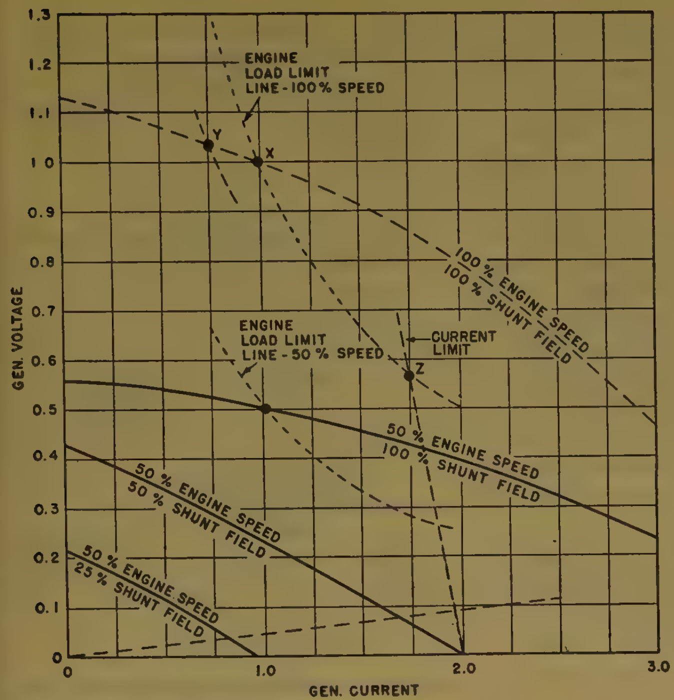
Suppose now the generator load has been increased slowly (or suddenly) until the "slave piston" has the throttle wide open and against the fuel stop. Suppose a further increment of load is imposed by the generator on the engine. The engine slows momentarily, unbalances the pilot valve and moves the "slave" piston. A spring between the "slave" piston and the fuel rack permits the former to move further.

This additional movement of the "slave" piston operates a rheostat R_3, which reduces the generator field, and therefore the load, just sufficiently to maintain engine speed.

Let us go back now to the instant where the load on the generator has just brought the throttle to full throttle position. If suddenly two cylinders cut out because of, for example, clogged fuel injectors, the engine will slow down momentarily and the "slave" piston will move. Since the fuel rack is already against the stop, the "slave" piston will simply reduce the generator field and thus keep the remaining cylinders fully loaded, but not overloaded. In this way the governor recognizes the engine's ability to take load. Refer to Figure 5 where this condition is illustrated by a generator curve.

With R_1 in the maximum engine speed position, the usual differential compound-wound generator volt-ampere characteristic for 100 per cent engine speed and 100 per cent field current is followed from no-load current up to point *X*. At this point the governor reaches the fuel stop, and the load limit rheostat R_3 shown in Figure 4 takes over and reduces the generator field. As more ampere load is put on the generator, R_3 makes the generator characteristic follow the load limit line shown in Figure 5, which is, of course, a constant kilowatt or horsepower load on the engine.

If a couple of cylinders were to cut out,



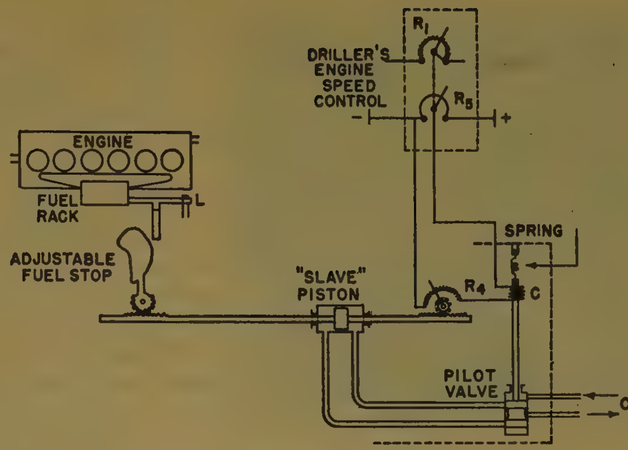


Figure 6 (above). Sketch of electrohydraulic engine governor showing mechanism for adjusting throttle stop as a function of engine speed

Figure 5 (left). Generator characteristics with electrohydraulic governor load limit and amplidyne current limit

the fuel stop would be reached earlier at some such point as Y, and the generator volt-ampere characteristic would follow a constant kilowatt or horsepower curve below the other. Thus the governor automatically adjusts the load to suit the engine condition. This feature is very valuable as the engines get older and less able to carry the same load than when new and in good repair.

Of course, as the driller's engine speed control rheostat R_1 is moved, the whole voltage characteristic moves up and down, as shown in Figure 5.

Throttle Stop Positioning. As yet nothing has been said about adjusting the position of the throttle stop as a function of engine speed. This adjustment is done with a completely separate mechanism from that so far discussed. As shown in Figure 6, the fuel stop is adjusted by a "slave" piston and solenoid-operated pilot valve which are, mechanically, duplicates of those used for the fuel-rack control. This system simply positions the "slave" cylinder in response only to the position of the potentiometer rheostat R_5.

Rheostat R_5 is on the same shaft as the driller's engine speed control rheostat R_1, and, together, they constitute the driller's speed master.

Note that the "slave" piston drives a cam-shaped stop for the fuel rack. The shape of this cam is determined by the engine manufacturer's curve of speed versus maximum-fuel, which is not a straight line.

Any kind of servomechanism would serve as well as that shown to position the fuel-stop cam as a function of engine-speed rheostat position, but a solenoid-controlled pilot valve and "slave" valve are most convenient, since oil under pressure is already available in the governor. The parts are duplicates of those on the throttle control.

All the solenoid valves, "slave" pistons, fuel stops, oil pressure regulating valve, and rheostats R_2, R_3, and R_4 are housed in one frame and mounted on the engine, convenient to the fuel rack, as illustrated in Figure 7.

The antihunt circuit for the governor shown in Figure 4 consists of rheostat R_2, a capacitor, and an AH coil on the pilot valve. A movement of the "slave" piston transiently introduces a force in the pilot valve in the opposite sense. This stabilizing action is very effective.

Advantages of Governor

To recapitulate, the advantages of the electrohydraulic Diesel engine governor for d-c drilling rig equipments are as follows:

1. It automatically prevents overloads on the engine, thus removing one of the principal causes of engine wear and failure.

2. It provides engine speed control by the driller at a remote location by means of a very simple rheostat and electrical connections. It thus permits the driller to run the engines at the lowest speed consistent with the required drilling operation. This feature promotes both fuel economy and low maintenance.

3. Paralleling several engine-generator sets, as required by the draw-works motor, and controlling them from one driller's controller is very simply accomplished by mechanically paralleling the speed control rheostats R_1. Proper division of load among the several engine sets is provided automatically by the governors.

4. The governor provides a "wrap-around" or constant kw characteristic of the generator around the engine characteristic, as shown in Figure 5. This characteristic permits the maximum torque at the maximum speed of the drilling motors at all points within the ability of the engines, instead of only at one fixed point as with conventional 2- or 3-field generator characteristics. This feature is, in effect, an automatic gear shift providing a continuously adjustable choice of ratios between the engine and the driven load.

Amplidyne Control System

An amplidyne is used, instead of a conventional exciter, to excite and control the main generator field because the greater amplification factor of the amplidyne requires only a few watts for the amplidyne field. This factor makes possible the very small and extremely sturdy rheostats used in the governor (illustrated in Figure 7) and in the drillers speed master.

The first requirement of the control system is that the generator voltage be manually controlled and adjusted in either direction. The principles of this system are shown in Figure 8.

The driller's control rheostat R_6 is a potentiometer across a source of d-c control power. A potential V_2 from the potentiometer R_6 is compared with a potential V_1 through the main amplidyne control field F_1F_2. Note that V_1 is a potential drop across a resistor A in the

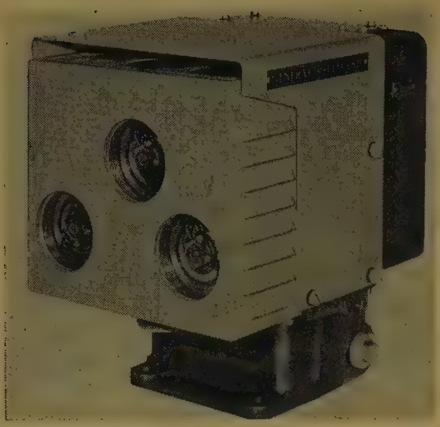


Figure 7. An electrohydraulic governor (tachometer not shown)

main generator field. Thus movement of the driller's rheostat R_6 adjusts the main generator field, and, as a result, the generator characteristics of Figure 5 for different values of generator field current are obtained. The very high amplification factor of the amplidyne makes V_1 always very nearly equal to V_2.

A set of forward and reverse contactors are actuated by pulling out the driller's master switch handle before rotating it, giving forward or reverse generator field excitation.

Instead of comparing the voltage V_2 with main generator field current, it might have been compared directly with the main generator voltage. However, if this were done, the amplidyne would try to nullify the action of the generator differential series field. Each time the main generator differential series field tried to pull the voltage down, the amplidyne would try to pull it up. Also, as has been proved by many types of installations, the stability of the system is much better if the control potential V_2 is balanced against main generator field current, leaving the differential series field to do its duty.

The engine load-limit potentiometer rheostat R_3 is inserted between the driller's control rheostat R_6 and the amplidyne control field F_1F_2, and is the same governor-actuated rheostat R_3 as shown in Figure 4.

Since several generators are to be controlled in parallel from one driller's control master switch, when hoisting, several amplidyne fields can be paralleled on one R_6 rheostat.

The potentiometer R_6, shown in Figure 8, and the engine speed control rheostat R_1 (Figure 4) are on one shaft and form the complete drillers speed master. The complete travel of the drillers speed master is about 330 degrees. The first 165 degrees of movement raises the generator field from zero to full value, but does not raise the engine speed. The next 165 degrees cut in R_1 to speed up the engine.

Current Limit

The next function to be provided is current limit so that the commutating ability of the generators and motors will not be exceeded, even if the motors stall. Likewise, it is desirable to limit the applied torque on the mechanical equipment to prevent damage. For example, the rotary-table-drive motor, when rotating a long string of drill stem, may require a definitely limited maximum torque to prevent twisting it off.

It should be understood clearly that the current limit function now under discussion is entirely separate and distinct from the engine load limit previously discussed. Current limit control is provided by a separate field on the amplidyne. This control field F_5F_6, illustrated in Figure 9, always is excited in the opposite direction from the control field F_1F_2 and must overpower it, and thus reduce main generator excitation whenever the main generator current reaches a certain value.

The current limit must be ineffective until a certain selected value of current is reached, regardless of the direction of flow of that current, and be effective when that value is reached.

In Figure 9 is shown the elementary diagram of the amplidyne connection used to accomplish this function.

A pair of rheostats R_7 are placed across a source of d-c control power. This reference voltage is compared, through a pair of blocking selenium rectifiers, to the drop across the commutating and series fields, which is proportional to generator current.

For one direction of main generator current, V_3 must be greater than V_4 before blocking rectifier BR_1 will permit current to flow in current limit field F_5F_6.

When the generator current reverses, which depends on a reversal of the control field F_1F_2, the drop V_5 is in the opposite direction. When V_5 exceeds V_6, the blocking rectifier BR_2 permits current to flow in the current limit field F_5F_6 in such a direction as to oppose control field F_1F_2.

The effect of current limit control is shown in Figure 5. The cut-off point Z can be adjusted for any desired value of current by adjusting rheostat R_7. While current limit begins to cut off at the control setting of point Z, the generator current has to rise a further 10 or 15 per cent

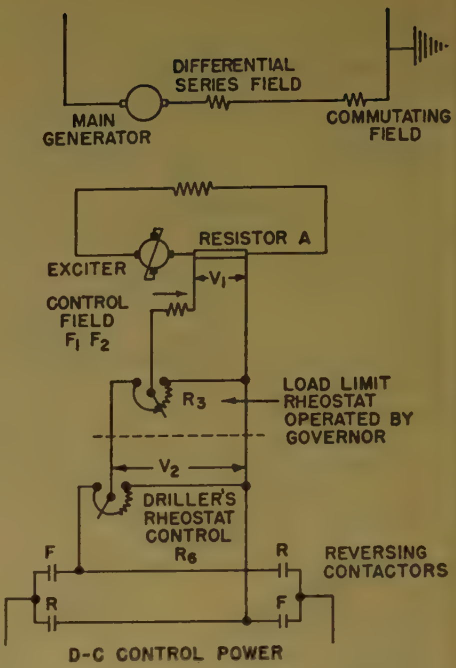


Figure 8. Elementary diagram of amplidyne field and engine load limit control

before the current limit field F_5F_6 can overcome completely the control field F_1F_2 and bring the generator voltage to zero.

Antihunt and Paralleling

An antihunt field F_7F_8 is provided on the amplidyne so that any sudden rise of amplidyne voltage causes a transient current to flow through the capacitor and control field F_7F_8, which opposes the change.

A paralleling field F_9F_{10} is connected, as shown, and connects to a similar arrangement on any additional generators connected in parallel. It tries to make the generator divide the loads and operates exactly by the same manner as a paralleling connection on regulators.

The paralleling field F_9F_{10} shown in Figure 9 is quite weak in comparison with the other fields and is mainly effective at light generator loads or for very sudden transients (for example, one generator out of several in parallel suddenly losing its field). Paralleling field F_9F_{10} is completely overcome and supplanted in function whenever the load limit feature of the governor goes into action.

Auxiliary Drives

In addition to the main drive motors, there are a number of auxiliary motors for driving ventilating blowers, pumps, air compressors, shale shakers, blowout preventer, and bug fans. The require-

ments for such motors may total 50 or 60 horsepower on a typical rig. Also, six or seven kilowatts is required for adequate lighting. Approximately 15 kilowatts of excitation power is required for the main drive motors.

On Diesel-electric rigs, it has been the practice in the past to add a constant potential d-c auxiliary generator to each engine. A typical rating is 30 kw, 125 volts d-c (125 volts has been selected in the past because of the lighting requirements).

Because the main engines are usually started by compressed air, a small battery-started gasoline engine-driven generator of five or ten kilowatts is installed to supply power to the air compressor motor, when the main engines are down, and to supply lighting power when rigging up or tearing down the rig.

This almost universal practice of supplying 125-volt d-c auxiliary power requires expensive d-c auxiliary motors and d-c control, generally with explosion-proof enclosures.

Since the auxiliary drives require only constant-speed motors, a-c auxiliary power is recommended. This a-c power permits very simple sturdy control and the auxiliary motors to be squirrel-cage induction motors, explosionproof if desired.

A very careful analysis of the costs of d-c versus a-c auxiliary power reveals that the cost of auxiliary equipment for a-c power is about 25 per cent cheaper than for d-c power. In making this comparison, two full capacity engine-driven 440-volt, 3-phase 60-cycle alternators were included. One of these alternators is a complete spare to the other, since the entire rig depends on this auxiliary power being available.

The cost comparison includes the cost of two auxiliary engines, but takes credit for eliminating the small 5-kw emergency lighting engine generator and reducing the size necessary for the main engines.

The use of entirely separate engine-driven auxiliary power generators is made even more desirable by the use of adjustable speed main engines. If d-c auxiliary power generators are driven by the main engines, they have to give full output at the lowest engine speed. This condition increases the cost of such d-c auxiliary equipment, and the foregoing cost comparison was made on the basis of running the main engines over a two to one speed range. With a-c auxiliaries, the lighting circuits are standard 115 volts, stepped down from 440 volts by transformers.

The advantages of a-c versus d-c

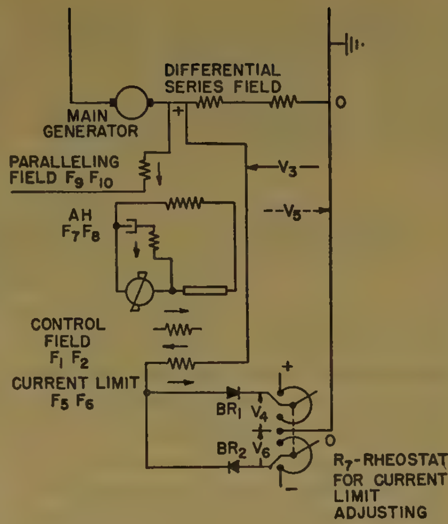


Figure 9. Elementary diagram of amplidyne control showing current limit, antihunt, and paralleling fields

auxiliaries can be summarized as follows:

1. About 25 per cent reduction in first cost.

2. The main engine-driven-generator units are simplified, lighter, easier to move and set up because of the elimination of the top-mounted belt-driven constant-voltage d-c auxiliary generators, and the consequent reduction in size of the main engine required.

3. The main control panel is reduced in size and weight because of simpler a-c instead of d-c control for auxiliaries.

4. Auxiliary a-c power at 440 volts, 3 phase, requires smaller cable to distribute power than does 125 volts d-c.

The disadvantages of having two small 100 or 125 horsepower internal combustion auxiliary engines to maintain are more than overcome by not having to run the main engines when only auxiliary power is required. Also, the 5- or 10-kw engine-driven set is eliminated.

Ventilation

The main rig motion motors are usually force-ventilated because they frequently are required to operate at very low speeds and full torque. Also, it is desirable to bring in fresh air, free from explosive gas mixtures, to eliminate the explosion hazard which would otherwise be present when the well strikes gas.

The most usual method of ventilation is to mount a small motor-driven blower on the top of each main d-c motor and run a duct away from the derrick to a safe source of fresh air.

These ducts have been made of steel in the past and are considered a nuisance to tear down and move approximately every two months. There is now a probability of obtaining flexible rubberized non-collapsible ducts which will be very easy to move.

The generators are self-ventilating and are located away from the derrick in a nonhazardous area.

On barge-mounted rigs, there is usually a central blower system, and air is piped to the various motors.

Because of the duty on the coring-reel motor, it usually can be totally enclosed and one-hour rated, thus eliminating the cumbersome duct work. However, to make this large totally-enclosed motor truly explosionproof would be very expensive, but it is a very simple matter to infiltrate a very small amount of safe air from the compressor system to keep this motor purged and under slight pressure.

Frequently, the mud pumps are located far enough away from the rig so that no duct work is necessary to bring in safe air, and their blowers take air directly from the ambient surroundings.

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Dynamic Braking Control of D-C Series Motors—Calculation of Stability Limit

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Synopsis: This paper presents an extension of the conventional method of calculating speed-torque curves for d-c series motors used with dynamic braking controllers, which permits determination of the stability limit with overhauling loads such as occur on crane hoists. It is shown that such stability limit can be increased by the addition of series resistance.

IN laying out control equipment, the design engineer looks for methods which enable him to calculate, with a minimum of effort and with reasonable accuracy, the speed-torque curves obtained on various controller points. Direct-current series motors are widely used for heavy duty intermittent drives, such as cranes, hoists, ore and coal unloaders, and other material handling equipment. Control of such motors is obtained by inserting resistance in the various branches of the motor circuit, that is, the armature and the field.

Considering that commercial resistors may deviate as much as ten per cent from their rated resistance, that line voltage may vary by ten per cent or even more, and that it is difficult on most motor applications to predict actual load conditions within a few per cent, it does not appear necessary to forecast motor performance with a greater accuracy than a few per cent. It is of greater importance to have a method available which permits making approximate calculations quickly so that control systems can be compared and analyzed without spending an undue amount of time and effort.

A simple method of calculation which is widely used by design and application engineers has been described by A. A. Merrill.[1] The information required from the motor designer is the characteristic curves of the motor, in which speed and torque are plotted as a function of motor current, as well as armature resistance $R_{a,b}$ commutating field resistance R_c, series field resistance R_f, and equivalent brush drop resistance R_b. As the armature and field current of a series motor are normally the same, all values entering into the calculations can be expressed in terms of either field or armature current.

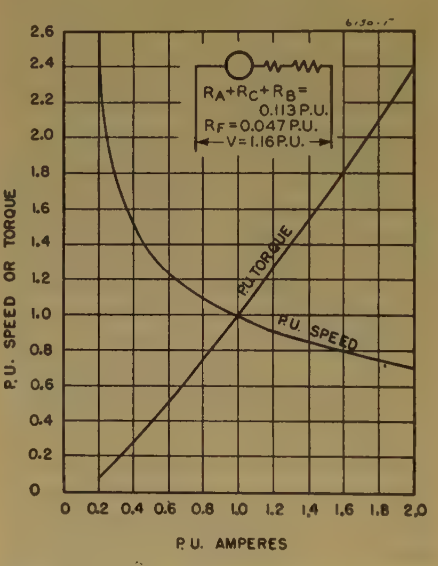

Figure 1. Per unit characteristic curves of d-c series motor

Average of line of 30-minute-rated mill motors

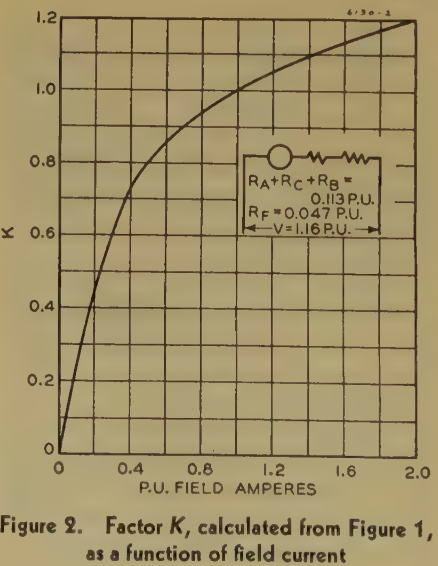

Figure 2. Factor *K*, calculated from Figure 1, as a function of field current

To obtain results which are generally applicable to motors of various sizes, it is convenient to express operating quantities in "per unit" (pu), rather than in actual numerical values. Rated or normal values are called base values, expressed as unity, and all other values are expressed as a decimal fraction of the base values. Base quantities used in this paper are defined as follows:

base current = rated full load current
base torque = rated full load torque
base speed = rated full load speed
base voltage = generated armature electromotive force at rated torque and rated speed
base resistance = base voltage divided by base current
base flux = flux per pole at rated full load speed and current

Figure 1 represents the average characteristic curves of a line of 30-minute-rated mill type d-c series motors. All values are expressed in per unit, based on rated 30-minute load. The actual load expressed by base current or base torque is higher than the same machine would be required to carry if it were rated on a continuous basis. Consequently, the magnetic and electric loadings are high, and the effects of saturation and armature reaction are more noticeable than for continuous rated machines of similar physical size, when loads are expressed in per unit, based on rated values.

Conventional speed-torque calculations are based on two fundamental equations for dynamo electric machines which state that

$$S=\frac{E}{k_s\phi} \tag{1}$$

and

$$T=k_t\phi I_a \tag{2}$$

where

E = generated voltage
T = developed torque
ϕ = magnetic flux per pole
I_a = armature current
S = speed of rotation
k_s and k_t = constants, the value of which is determined by motor design and by the units of E, R, I_a, and S

Let

$k_s\phi=K_s$ and $k_t\phi=K_t$

Then

$$S=\frac{E}{K_s} \tag{3}$$

and

$$T=K_tI_a \tag{4}$$

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The authors wish to acknowledge the assistance given them in the preparation of this paper by C. W. Lange, engineer, d-c armored motor engineering division, General Electric Company, who first called attention to the effect of series resistance on the stability limit.

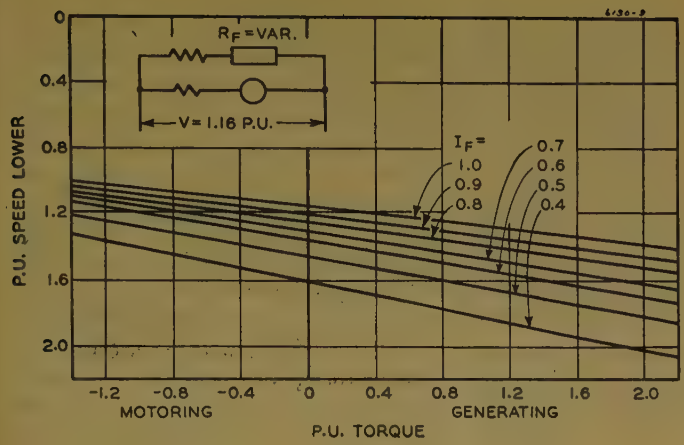

Figure 3. Speed-torque curves of a dynamic braking controller, calculated with constant K

In the per unit system, $K_{s(pu)}$ and $K_{t(pu)}$ have the same numerical value because they are both directly proportional to the magnitude of the flux. This may be shown as follows:

$$K_{s\,base} = \frac{E_{base}}{S_{base}} = k_s\phi_{base}$$

$$K_{s1} = k_s\phi_1$$

$$K_{s(pu)} = \frac{K_{s1}}{K_{s\,base}} = \frac{k_s\phi_1}{k_s\phi_{base}} = \frac{\phi_1}{\phi_{base}}$$

$$K_{t\,base} = \frac{T_{base}}{I_{a\,base}} = k_t\phi_{base}$$

$$K_{t1} = k_t\phi_1$$

$$K_{t(pu)} = \frac{K_{t1}}{K_{t\,base}} = \frac{k_t\phi_1}{k_t\phi_{base}} = \frac{\phi_1}{\phi_{base}}$$

Therefore

$$K_{s(pu)} = K_{t(pu)} = K \qquad (5)$$

Also

$$K_{s(pu)} = \frac{E_{(pu)}}{S_{(pu)}} \qquad (6)$$

and

$$K_{t(pu)} = \frac{T_{(pu)}}{I_{a(pu)}} \qquad (7)$$

Therefore:

$$K = \frac{E_{(pu)}}{S_{(pu)}} \qquad (8)$$

or

$$K = \frac{T_{(pu)}}{I_{a(pu)}} \qquad (9)$$

and

$$S_{(pu)} = \frac{E_{(pu)}}{K} \qquad (10)$$

$$T_{(pu)} = KI_{a(pu)} \qquad (11)$$

Values of K may be calculated from Figure 1 by equation 8 or 9. When equation 8 is used, $E_{(pu)}$ is obtained from the equation

$$E_{(pu)} = V_{(pu)} - I_{a(pu)}R_{a(pu)} \qquad (12)$$

where

$V_{(pu)}$ = per unit line voltage
$I_{a(pu)}$ = per unit amperes in armature circuit
$R_{a(pu)}$ = per unit ohms in armature circuit

For generator operation, $E_{(pu)}$ is given by

$$E_{(pu)} = V_{(pu)} + I_{a(pu)}R_{a(pu)} \qquad (13)$$

The data so calculated are then plotted against field amperes as illustrated by Figure 2. The data shown in Figure 2 are based on values and curves given in Figure 1.

Speed and torque for any given set of conditions can be calculated using Figure 2. When field current, armature current, line voltage, and motor resistance are known, generated voltage is calculated by equation 12 for motor operation, or equation 13 for generator operation. Speed is then obtained from equation 10, and developed torque from equation 11, using the value of K corresponding to the given field current. The shaft torque may be obtained by multiplying the developed torque by torque efficiency for motor operation, or dividing the developed torque by torque efficiency for generator operation. In subsequent speed-torque calculations, torque efficiency is considered to be unity.

A family of speed-torque curves for a dynamic braking controller, calculated by the method described above, is given in Figure 3. In this type of controller, the series motor is connected as a shunt machine on lowering points. Heavy loads overhaul the motor, which acts as a shunt generator and develops a dynamic braking torque restraining the load. This torque is indicated as positive in Figure 3. An empty hook or a light load is driven down by the motor, which acts as a shunt motor, deliverying a torque to accelerate the load. This torque is indicated as negative. Line voltage used for the calculation is 1.16, total internal motor resistance being 0.16.

The calculated curves of Figure 3 show that any desired no-load speed can be obtained by inserting sufficient resistance in series with the field. The straight-line curves give the impression that speed varies only slightly within the operating range between motoring and generating. This conclusion is incorrect. J. A. Jackson[2] has called attention to the fact that when operating with weakened field, the motor develops less dynamic braking torque than the curves of Figure 3 indicate. As braking torque and armature current increase, armature reaction plays an important role, reducing the effective flux. For a given overhauling torque, the speeds are higher than the calculation indicates. There is a definite limit of dynamic braking torque which the machine can develop with a given

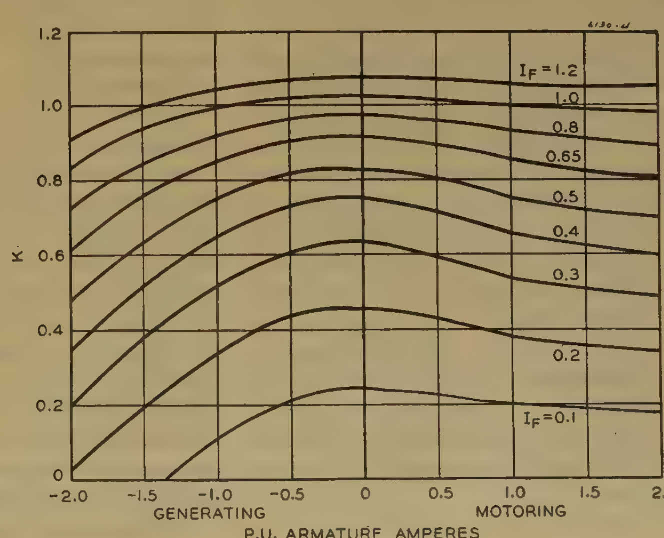

Figure 4. Factor K for d-c machine operating as motor or generator

Average of line of 30-minute-rated mill motors

amount of field current. If the load exceeds this torque, the machine becomes unstable, and a runaway condition results. To calculate the actual performance and the stability limit, it is necessary to abandon the concept that K is an independent function of field current.

Stability Limit

To obtain a true picture of the variation in the value of K with variation in armature current, it is necessary to test the motor over a range of loads both motoring and generating at several different values of field current. The average results, expressed in per unit values, of a number of such tests on a line of mill type d-c series motors are given in Figure 4. It has already been shown that the value of K is directly proportional to the magnetic flux in the motor, and it is evident from the curves of Figure 4 that the magnitude of the armature current does materially influence the amount of magnetic flux. This influence is the result of two factors:

1. Demagnetization produced by cross-magnetizing armature reaction.
2. Magnetization or demagnetization produced by currents flowing in armature coils undergoing commutation.

The theory of demagnetization produced by armature reaction may be found in any textbook on d-c machinery. In brief, the cross-magnetizing reaction produces a distortion in the flux distribution between the main poles and the armature which, because of the characteristic shape of the saturation curve, results in a net loss of flux. The effect is the same whether the machine is operating as a motor or as a generator.

The magnetizing or demagnetizing effect produced by current in the armature coils undergoing commutation would not be noticeable in a commutating pole machine if compensation could be adjusted exactly for all conditions of load. However, the magnetic loading of 30-minute-rated d-c motors is high, and this leads to saturation of the commutating pole magnetic circuit at heavy loads. When saturation occurs, the machine is operating in an undercompensated condition, and the current in the armature coils short-circuited by the brushes then has sufficient magnitude and such direction as to produce a direct magnetizing or demagnetizing effect on the main pole magnetic circuit. Analysis shows that this effect is demagnetizing for generator operation and magnetizing for motor operation.

Considering the combined effect of

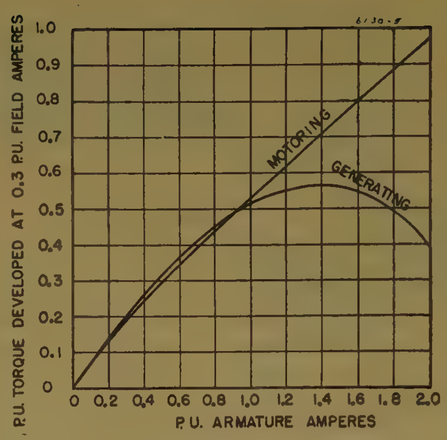

Figure 5. Developed torque, motoring and generating

these two phenomena, it will be noted that they are additive and demagnetizing for generator operation. It is to be expected that the result would be an appreciable loss of flux, and the test results presented in Figure 4 confirm this expectation. For motor operation the two effects are opposing, and the loss of flux at heavy overloads is much less, particularly at very heavy and very light field currents where the effect of cross-magnetizing armature reaction is minimum.

The loss of flux with increasing armature current when operating as a generator results in very definite torque limitations. Equation 11 states that developed torque is equal to the product of K and armature current. Since K is an inverse function of armature current for generator operation, the product has a definite maximum value, as illustrated by Figure 5. It is the maximum value of torque which determines the limit of stable operation. When it is exceeded, even under transient conditions, dynamic braking control is lost and cannot be regained, because increase in speed corresponds to decrease in torque. With loss of braking control, the load quickly accelerates to a dangerous speed.

In order to indicate the significance of variable K to the speed-torque curves of a dynamic braking controller, Figure 6 shows the actual performance curves obtained with a series motor, connected as a shunt machine. These curves apply to the same circuit and field conditions as Figure 3, but variable K has been used from Figure 4. The no-load speeds obtained by the two methods of calculation differ only slightly. In the region of power torque the curves differ somewhat, and especially with weak field the slope of the curves with variable K is less than for the curves calculated with constant K.

Performance in the dynamic braking region shows the most significant difference between the two methods of calculation. Figure 6 indicates that, with heavy loads, considerably higher speeds are obtained than would be expected from Figure 3. Also, for each value of field current, there is a definite torque beyond which the machine is unable to retard an overhauling load, that is, beyond which the machine becomes unstable. Assuming that an average crane hoist has a mechanical efficiency of 0.85, a dynamic braking torque of $(0.85)^2$, or 0.72, would then be required to lower full rated load. This torque is dangerously close to the motor stability limit with a field current of 0.5 or less.

Speed-torque curves such as plotted in Figure 6 represent steady-state operating conditions. While switching between controller points, transient armature current peaks and corresponding torque peaks occur. Even though the steady-state stability limit may be greater than full-load torque, switching transients may cause the stability limit to be exceeded, resulting in a runaway condition. This means that the control designer is not free to select any no-load speed he can obtain by field weakening. The amount of field weakening which is permissible, and thus the no-load speed which can be realized, are definitely limited by motor stability when lowering full load.

Effect of Series Resistance

A simple method of increasing the stability limit is to insert a block of resistance in series with the parallel circuit formed by the armature and the field. Resistance between 0.10 and 0.20 has only a slight effect on no-load speed and on the slope of the curves in the motoring quadrant. In Figure 7 a set of speed-torque curves is plotted for the same circuit conditions as in Figure 6, except that series resistance of 0.15 has been added. For each curve the amount of resistance in series with the field is held constant at such a value that the indicated value of I_f would be obtained if R_s were zero. On each curve there is a point at which no current is taken from the line, and at this point the indicated value of I_f actually flows through the field, whereas under all other operating conditions field current is either somewhat smaller or larger.

Series resistance has the effect of changing the distribution of current between field and armature at any given operating point. In the range of motoring and light overhauling torques, field current is smaller than it would be without series resistance. This is of no practical conse-

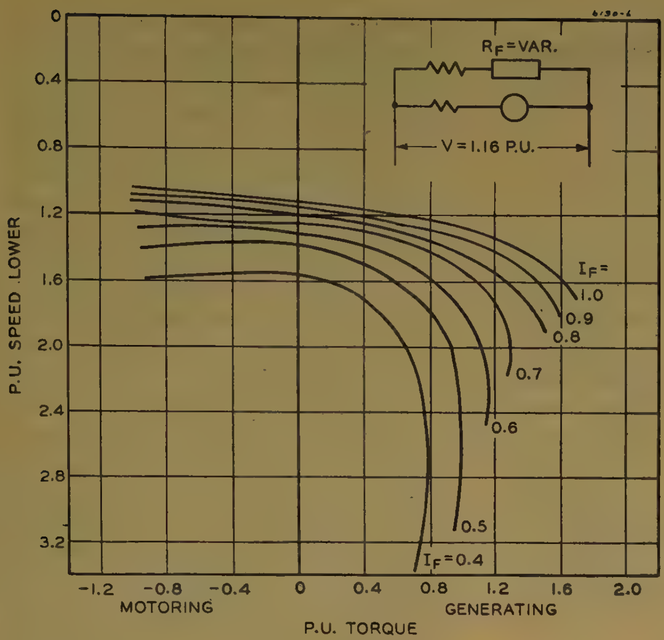

Figure 6. Speed-torque curves of a dynamic braking controller, calculated with variable *K* from Figure 4

No series resistance

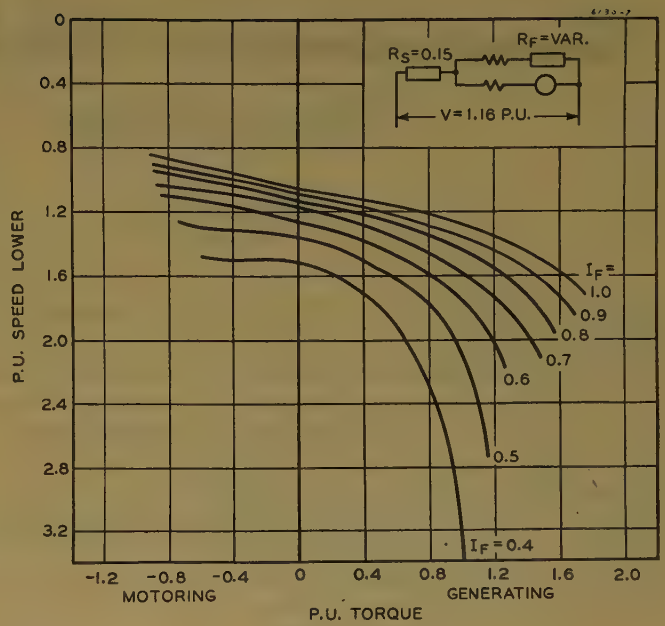

Figure 7. Speed-torque curves of a dynamic braking controller, calculated with variable *K* from Figure 4

With series resistance

Values of field current I_F indicated on curves exist only for the condition that line current is zero. Field resistance R_F for each curve is the same as in Figure 6

quence, as this is a stable motor operating range. Under high overhauling load conditions, however, the field current is greater. As factor K increases correspondingly, the motor is able to develop a higher braking torque, and the stability limit occurs at higher load. Comparing Figure 7 with Figure 6 for each value of field current, the slope of the curve is decreased in the higher torque region.

Assuming that the controller design permits transient torque peaks of twice normal when switching between controller points, the stability limit should occur at a braking torque of not less than twice the torque required to lower rated full load on the hook, that is, $2 \times 0.72 = 1.44$. Without series resistance, a field current of 0.70 would result in a runaway. With series resistance, a field current of 0.60 (at zero line current) would still be permissible.

Conclusions

The method of calculating speed-torque curves by using a variable factor K can be used for any d-c series motor for which sufficient test data are available for determining K both for motoring and generating operation. While the per unit data given in this paper permit the calculation of performance curves for the purpose of comparing various control circuit designs, actual motor test data may vary over a considerable range and must be used to determine the performance of any given motor application.

In designing dynamic braking controllers, using the motor field with suitable resistance in parallel with the armature, resistance in series with the parallel circuit formed by field and armature should be used to increase the motor stability limit.

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Operating Experience With Distance Ground Relays

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DISTANCE RELAYS for protection against wire-to-wire, two wire-to-ground, and three phase faults on transmission lines have met with wide acceptance. However, the use of distance relays for protection against single wire-to-ground faults has been limited to relatively few installations. From the information available, it would seem that the Kansas Gas and Electric Company is one of the few companies using distance ground relays. The experience gained in the use of these relays over the past seven years has demonstrated their reliability and good operating performance.

Theory of Distance Ground Measurement

The theory of distance ground measurement was first presented in 1931.[1] In general, distance relays compare the magnitude of a voltage with that of a current, the ratio being an indication of the distance to the fault. Zero sequence current and voltage (usually associated with ground relaying) are not suitable for distance ground relaying, because a comparison of these quantities is not a measure of the distance to the fault. The zero sequence voltage at the relay is equal to the sum of the zero sequence voltage drops from the relay to the source of ground current, and not from the relay to the fault as is required in distance relaying. Instead of these quantities, distance relays for ground faults use phase-to-ground voltage with the current in the corresponding phase, thus requiring three relays for the protection of each 3-phase line. This number is in addition to the relays required for protection against phase faults. For correct distance measurement under all system conditions, reactance type relays must be used and suitable compensation must be applied. The compensation used involves the zero-sequence current in the line section being protected and in the parallel line, if any.

As already shown,[1,2] the general expression for the line-to-ground voltage at the relay during a line-to-ground fault (see Figure 1) is

$$E_{aG}=mZ_{1GH}\left(I_{aGF}+I_{0GF}\frac{Z_{0GH}-Z_{1GH}}{Z_{1GH}}+I_{0GH}\frac{M_{0GH}}{Z_{1GH}}\right)+(I_{aGF}+I_{aHF})\,R_G \qquad (1)$$

where

E_{aG} = line-to-ground voltage, A phase at G
m = fraction of distance from G to H at which fault is assumed to occur
I_{aGF} = phase A fault current in faulted line at G
I_{0GF} = zero-sequence fault current in faulted line at G
Z_{1GH} = positive-sequence impedance, station G to H
Z_{0GH} = zero-sequence impedance, station G to H
M_{0GH} = zero-sequence mutual impedance between the two parallel lines from G to H
I_{0GH} = zero-sequence fault current in parallel line
I_{aHF} = phase A fault current from opposite end of faulted line
R_G = fault resistance
G = substation where relaying is considered
H = substation at opposite end of line section
F = fault location

Looking at equation 1, it will be observed that if we could obtain a single current for the relay equal to the current expressed within the first brackets, namely

$$I_{\text{relay}}=I_{aGF}+I_{0GF}\frac{Z_{0GH}+Z_{1GH}}{Z_{1GH}}+I_{0GH}\frac{M_{0GH}}{Z_{1GH}} \qquad (2)$$

we obtain a very simple result, as follows:

$$Z_R=\frac{E_{aG}}{I_{\text{relay}}}=mZ_{1GH}+\left(\frac{I_{aGF}+I_{aHF}}{I_{\text{relay}}}\right)R_G \qquad (3)$$

where:

Z_R = impedance indicated by the relay
I_{relay} = current through relay

The first term of this expression is directly proportional to the distance from the fault to the relay and is independent of conditions external to the section being protected. The second term is a function of fault resistance R_G and, therefore, represents a voltage in phase with the fault current. By using a reactance type relay, which is responsive only to the reactive component of the voltage, the effect of the last term is rejected. This rejection is necessary because this last term varies with changing system conditions, and also because R_G itself is likely to vary over wide limits as compared with the fault resistance experienced in wire-to-wire faults.

In practice the proper relay current is obtained by the use of an auxiliary current transformer which takes the proper proportion of the residual current of the protected line and combines it with the proper proportion of the residual current from the parallel line (if any). In one type of distance ground relay, this current is added to the phase current I_{aGF} (equation 2), and the total fed into the relay. In another type of distance ground relay, the phase current I_{aGF} is fed into one winding of the relay, and the output of the auxiliary current transformer is fed into a second winding of the relay. Both windings then operate on the same element. The net result is the same.

A vector diagram of the currents and voltages involved in distance ground relaying is shown in Figure 2. In one type of distance ground relay, instead of using the voltage 90 degrees from the faulted phase for polarizing the directional element, as shown, residual or zero phase sequence current and voltage is used in the directional element.

Figures 5 and 6 show examples of typical installations of distance phase and ground relays.

Figure 7 is a typical wiring diagram for an installation of one type of distance ground relays. The phase relays are not shown.

Reasons for Use

There are several reasons why distance ground relays were considered justified and installed on the Kansas Gas and Electric Company system instead of ordinary directional ground relays or carrier pilot relays. The reason why Peterson coils could not be used also is explained.

Distance ground relays give the same high speed protection against ground faults that distance phase relays give against phase faults. Each line section is protected selectively without cascading settings. Although, in general, it is not necessary to cascade the settings of directional ground relays to anything like

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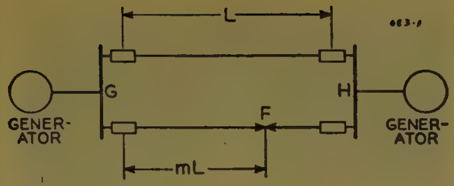

Figure 1. One-line diagram of assumed system

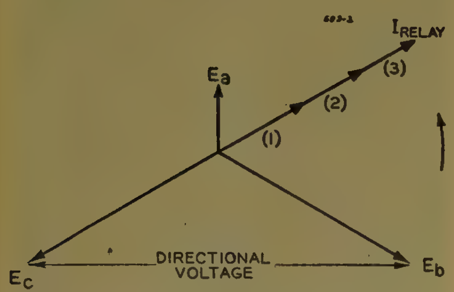

Figure 2. Vector diagram showing relation of currents and voltages for a one-wire-to-ground fault on phase *A*; I_{relay}=total of three components, namely:

(1). I_{aGF}=current in phase *A*

(2). $I_{0GF}\dfrac{Z_{0GH}-Z_{1GH}}{Z_{1GH}}$=a percentage of the zero-sequence current in the protected line

(3). $I_{0GH}\dfrac{M_{0GH}}{Z_{1GH}}$=a percentage of the zero-sequence current in the parallel line

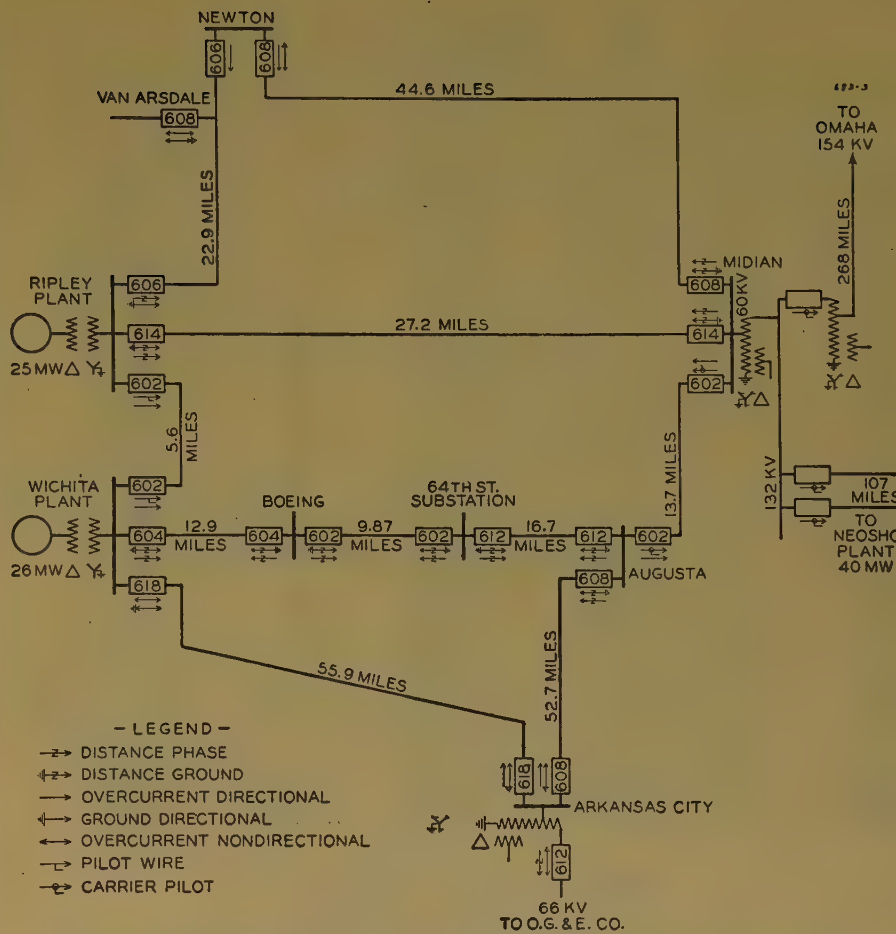

Figure 3. Central Kansas 60-kv system of the Kansas Gas and Electric Company

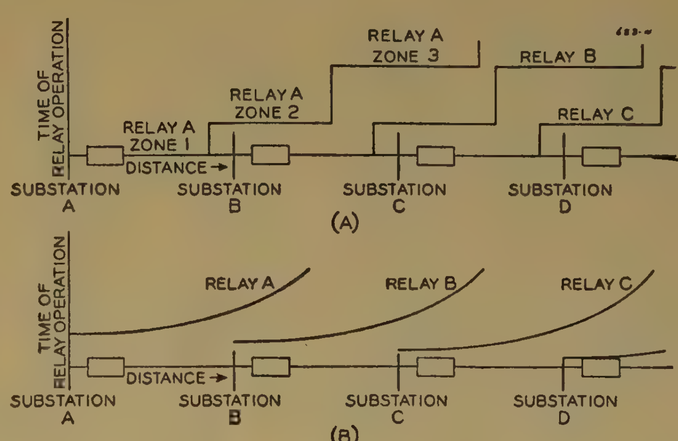

Figure 4

A—Distance relay characteristic; all but last 10 or 15 per cent of line is instantaneous trip

B—Induction relay characteristic

the extent necessary in the case of conventional overload phase relays, high speed distance ground relays do offer a very substantial saving in relaying time in most cases. This fact is especially true where there are a number of stations between grounding points, as on the loop between Wichita Plant and Midian, by way of 64th Street substation and Augusta (see Figure 3). The time saving to be gained by the use of distance relays instead of induction type ground relays is illustrated in Figure 4.

A very high percentage of all line faults on the Kansas Gas and Electric Company's 60-kv system are one wire-to-ground. For example, out of 451 "trip-outs," which have occurred on circuit breakers relayed with distance ground relays since their installation, 378 (83.8 per cent) were one wire-to-ground and 73 involved more than one wire.

Although the shock to the system is less severe for single wire-to-ground faults than for faults involving more than one wire, the large percentage of ground faults was considered justification for an expenditure for ground relaying equal to that for phase relaying.

Ground faults, when not cleared rapidly, frequently result in burned down conductors or overhead ground wires.

The cost of an installation of phase and ground distance relays is approximately half the cost of a carrier pilot relay installation. This cost makes complete distance relay protection for all types of faults very attractive in those cases where good relaying is desired, but where there exists no real necessity for going to carrier relaying.

It was not feasible to consider the use of Peterson coils on this system because of its complexity and number of grounding points. But, the main reason was that at Midian and Arkansas City (Figure 3) large autotransformers are used having graded insulation and solidly grounded neutrals which could not be isolated from ground by a Peterson coil.

Disadvantages

Distance ground relays have the disadvantage of being more complicated than distance phase relays. They require a set of auxiliary blocking relays, which block operation upon the occurrence of wire-to-wire or two wire-to-ground faults, in order to prevent overreaching into the next line section on these types of faults. As pre-

Table I. Typical Relay Operating Record

Ripley Midian Line

OCB Number		Date	Time	Relays	Cause
Ripley	614	3-15-44	11:42 a.m.	GB1	Lightning
Midian	614	3-15-44	11:42 a.m.	GB1	Lightning
Ripley	614	4-10-44	9:57 p.m.	GB1	Lightning
Midian	614	4-10-44	9:57 p.m.	GB1	Lightning
Ripley	614	4-22-44	1:46 a.m.	GB1	Lightning
Midian	614	4-22-44	1:46 a.m.	?	Lightning
Ripley	614	4-23-44	3:38 p.m.	GA1-A1	Lightning
Midian	614	4-24-44	3:38 p.m.	GA2	Lightning
Ripley	614	5- 1-44	1:02 a.m.	GB1	Lightning
Midian	614	5- 1-44	1:02 a.m.	GB1	Lightning
Ripley	614	5-29-44	4:39 p.m.	GB1	Lightning
Midian	614	5-29-44	4:39 p.m.	GB1	Lightning
Ripley	614	7- 8-44	9:17 p.m.	GB1	Lightning
Midian	614	7- 8-44	9:17 p.m.	GB1	Lightning
Ripley	614	7-10-44	3:17 p.m.	GB1	Lightning
Midian	614	7-10-44	3:17 p.m.	GB1	Lightning
Ripley	614	7-11-44	4:22 a.m.	GB1	Lightning
Midian	614	7-11-44	4:22 a.m.	GB1	Lightning
Ripley	614	7-18-44	8:42 a.m.	GB1-A1	Lightning
Midian	614	7-18-44	8:42 a.m.	GB2	Lightning
Ripley	614	7-25-44	3:03 a.m.	GB1	Lightning
Midian	614	7-25-44	3:03 a.m.	GB1	Lightning
Ripley	614	7-27-44	9:43 p.m.	GB1	Lightning
Midian	614	7-27-44	9:43 p.m.	GB1	Lightning
Ripley	614	8-13-44	5:50 p.m.	GB1-A1	Crossarm broken
Midian	614	8-13-44	5:50 p.m.	GB2	Crossarm broken
Midian	614	8-13-44	5:51 p.m.	GB2	Crossarm broken
Ripley	614	8-13-44	6:21 p.m.	GB1-A1	Crossarm broken
Ripley	614	9-10-44	4:20 a.m.	GB1	Lightning
Midian	614	9-10-44	4:20 a.m.	GB1	Lightning
Ripley	614	11- 2-44	10:37 p.m.	GB1	Lightning
Midian	614	11- 2-44	10:37 p.m.	GB1	Lightning

Table II. Typical Relay Operating Record

Augusta–64th Street Line

OCB Number		Date	Time	Relays	Cause
Augusta	612	1-27-44	12:28 a.m.	A1	Wind storm
64th Street	612	1-27-44	12:28 a.m.	A1	Wind storm
Augusta	612	4-26-44	12:37 a.m.	GC1	Lightning
64th Street	612	4-26-44	12:37 a.m.	GC1	Lightning
Augusta	612	6- 8-44	3:24 p.m.	GA1	Lightning
64th Street	612	6- 8-44	3:24 p.m.	GA1	Lightning
Augusta	612	6-18-44	7:25 p.m.	GB1	Lightning
64th Street	612	6-18-44	7:25 p.m.	GB1	Lightning shattered insulator
Augusta	612	6-18-44	7:26 p.m.	GB1	Lightning shattered insulator
Augusta	612	6-18-44	7:34 p.m.	GB1	Lightning shattered insulator
Augusta	612	7-27-44	9:53 p.m.	GC1	Lightning
64th Street	612	7-27-44	9:53 p.m.	GC1	Lightning
Augusta	612	8-24-44	7:01 a.m.	GB1	Lightning
64th Street	612	8-24-44	7:01 a.m.	GB1	Lightning

viously mentioned, they also require an auxiliary current transformer for zero-phase-sequence compensation. The experience has been that these disadvantages are not of such a nature that they should preclude the use of distance ground relays.

The relay installation for each oil circuit breaker consists of three distance phase relays and three distance ground relays, a total of six relays. It would be possible, by means of current and potential switching circuits, to reduce the number of relays required to one phase relay and one ground relay,[3,4] but with some increase in tripping time. It is believed that the scheme employing two relays has its principal application on subtransmission lines, 33 kv and lower, rather than on important 60-kv and higher voltage lines.

Relay Installations

The first installation on the Kansas Gas and Electric Company 60-kv system was completed October 6, 1938, and was followed from time to time by other installations, until at the present time ten installations of distance ground relays are in service.

A 1-line diagram of the Kansas Gas and Electric Company Central Kansas 60-kv system is shown in Figure 3. The loca-

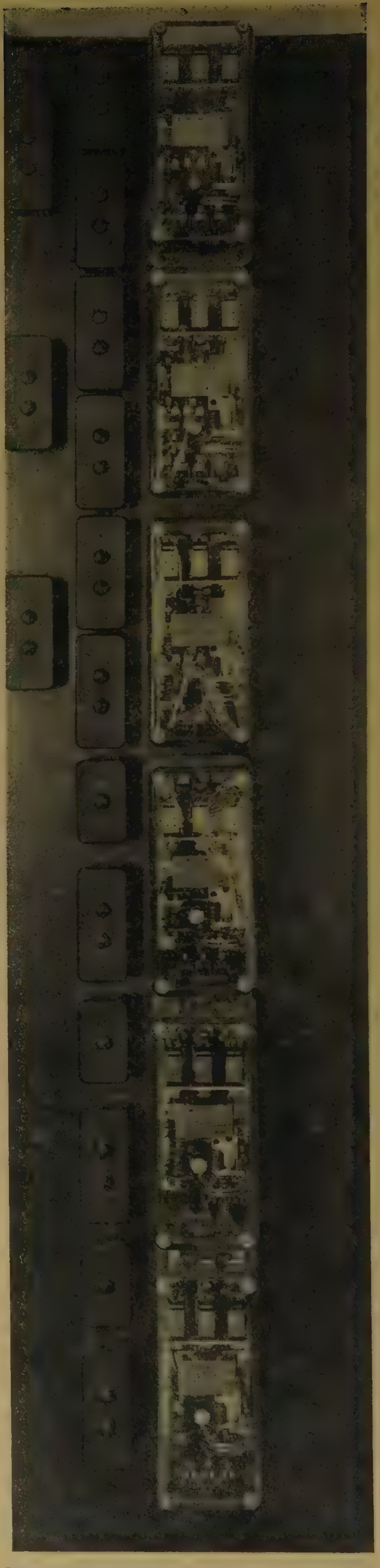

Figure 5. Installation of distance phase and distance ground relays (the three top relays are ground relays)

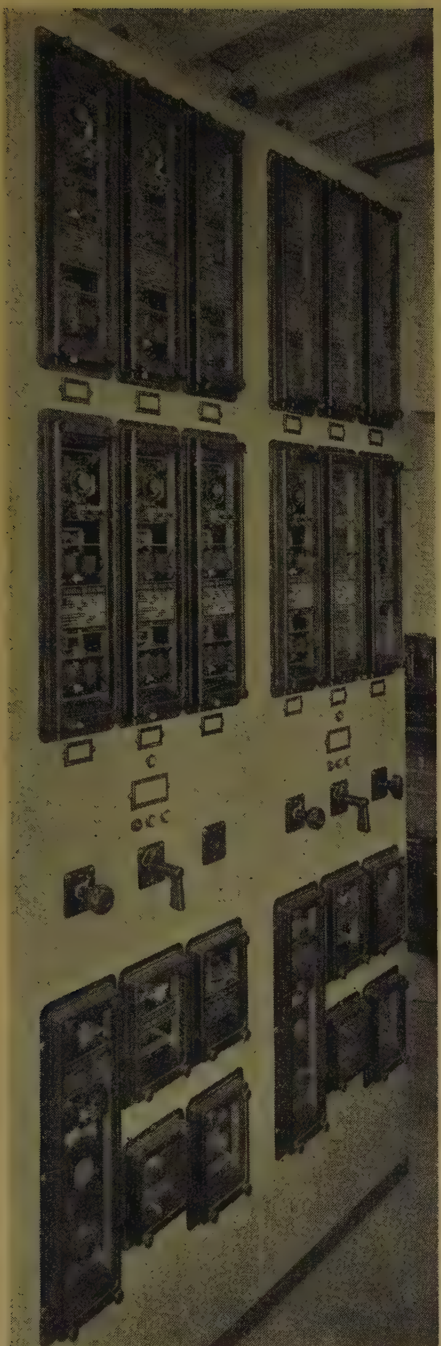

Figure 6. Installation of distance phase and distance ground relays for two 60-kv circuit breakers

The phase relays are across the top of the panel and the ground relays are across the middle. The relays in the lower section are, with the exception of the auxiliary blocking relay, associated with the ground relays for automatic reclosing and synchronous check functions

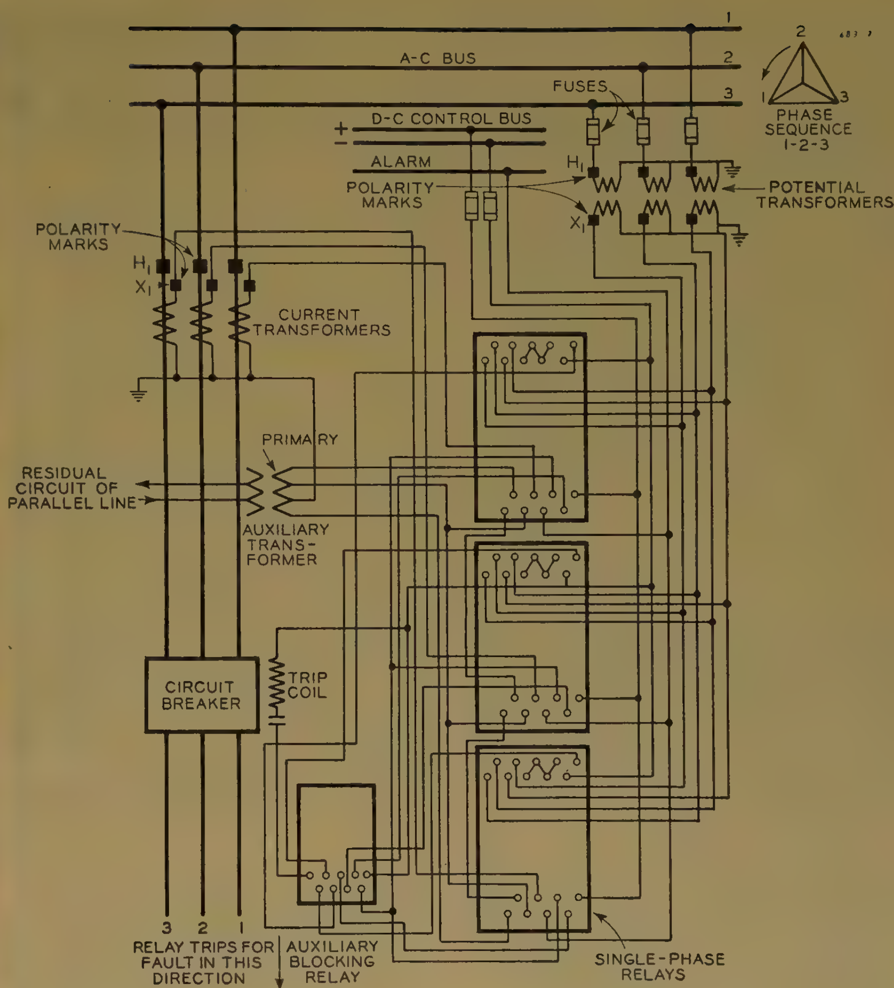

Figure 7. Typical wiring diagram for one type of distance ground relay installation

tion of distance ground relays is indicated by symbol on this diagram. The dates on which the various installations were completed are as follows:

Ripley 614..............October 6, 1938
64th Street 602..........April 4, 1941
64th Street 612..........April 4, 1941
Augusta 612.............May 9, 1941
Wichita 604.............June 7, 1941
Midian 608..............July 14, 1942
Midian 614..............July 23, 1942
Augusta 608.............August 17, 1942
Boeing 602..............February 4, 1943
Boeing 604..............February 4, 1943

Previous to the installation of distance relays and other modern relays on this system, induction overload and directional relays were used. Ground relays were installed on only a few of the circuit breakers. Co-ordination between the relays on the various circuit breakers was by time sequence, taking advantage of the diversity of current flow in the various circuits under fault conditions as much as possible. Instantaneous overload relays, set to cover 90 per cent of the line length under maximum generating conditions, were used on lines which were long enough to secure reasonable coverage under average generating conditions. At this stage, Boeing and 64th Street substations had not been built and the Wichita–Augusta line was one continuous line. Nor had Ripley been built, but there was a 60-kv substation not far from the plant site having three 60-kv circuit breakers, making the same system connections as the three circuit breakers at Ripley do now. Even with the help of the instantaneous relays on the longer lines, the relay time on the majority of the circuit breakers at Wichita Plant and Midian were quite long. In addition, it was found impossible to find relay settings which would fit all system conditions. Instability troubles frequently developed between Wichita and Midian because of the long relaying times involved.

Sometimes the opinion is expressed that zone type distance relays cannot be applied to a system on which they must be co-ordinated with induction overload relays. It is believed, however, that this fear has been exaggerated. No particular difficulty has been experienced in applying distance relays to the system under consideration. The distance relays were put in service, one or two installations at a time, over a period of several years, with the old induction relays continuing in service on the remaining circuit breakers. In fact the transition is still not complete since, as can be seen on Figure 3, six circuit breakers remain which have the

Table III. Standard Code for Marking Relay Targets

Kind of Protection	Function or Type	Symbol
Ground	Overload, time delay	G
Ground	Overload, instantaneous	G1
Ground	Distance; HX, GCX	GA1, GB1, GC1 (for zone 1); GA2, GB2, GC2 (for zone 2); GA3, GB3, GC3 (for zone 3)
Phase	Overload, time delay; IA, IAC, CO, CR, IB, and so forth	A, B, C
Phase	Overload, instantaneous	A1, B1, C1
Phase	Distance; HZ, GCX	A1, B1, C1 (for zone 1); A2, B2, C2 (for zone 2); A3, B3, C3 (for zone 3)

old relays. On most of these circuit breakers, instantaneous overload attachments help greatly in effecting good co-ordination, mostly by reducing the second zone setting which otherwise would be necessary on some of the distance relays.

Operating Results

Operating experience with distance ground relays has been excellent. They are called upon to operate from three to five times more than are the phase relays, and they can be depended on to operate correctly when properly applied.

An example of the operating performance of these relays is shown in Tables I and II, which give a part of the relay operating record of the Ripley–Midian line and the 64th Street–Augusta line respectively.

For the purpose of assisting in the understanding of the relay operations indicated in Tables I and II, Table III is included, which gives a partial list of the standard method of marking and reporting relay targets used by the Kansas Gas and Electric Company.

An analysis of the distance ground relay operations from October 1938 through 1945 discloses that there have been a total of 391 operations, of which 378 were correct in every respect. During the same period, there were a total of 75 distance phase relay operations (counting only those breakers on which distance ground relays also were installed.) Of these, 73 were correct in every respect. The summary of these operations is given in Table IV. Several instances of the first zone overreaching the next substation developed in the case of two installations, but this condition was corrected easily by shortening the distance setting slightly. The experience has been that if the relays are set to reach 85 per cent of the distance to the next station, there will be no trouble from overreaching. When set for 90 per cent, as is the usual practice for phase relays, the two instances mentioned occurred.

Table IV. Summary of Distance Relay Operations

Year	Ground Relays: Correct	Ground Relays: Questionable	Ground Relays: Wrong	Phase Relays: Correct	Phase Relays: Questionable	Phase Relays: Wrong
1938	2	0	0	0	0	0
1939	21	0	0	2	0	0
1940	9	2	1	1	0	0
1941	40	0	1	7	0	0
1942	40	0	2	21	0	0
1943	87	0	4	16	0	2
1944	108	0	3	21	0	0
1945	71	0	0	5	0	0
	378	2	11	73	0	2

This analysis may be summarized as follows:

Questionable Operations: Ground

1. Zone 2 fault tripped in zone 3 (or was so reported) 1
2. Two ground relay targets reported for one "tripout" (GA1 & GB1). Only one relay should operate because of blocking units 1

Total 2

Wrong Operations: Ground

1. Overreached. Tripped in zone 1 for fault beyond next substation 3
2. Bad auxiliary contact caused unnecessary "tripout" 2
3. Loss of potential (fuse blown) 6

Total 11

Wrong Operations: Phase

1. Failure to operate because of faulty contact adjustment 2

Total 2

Loss of potential was the cause of more false operations than any other one thing. It is expected that less trouble will be experienced from this cause in the future since heavier potential transformer fuses now are used, which are not likely to fail from deterioration. Instantaneous overcurrent fault detector relays can be used to prevent tripping from loss of potential. In order to keep the burden on the current transformers to a minimum, three such relays, mounted separately, should be used to supervise both the phase and ground relays rather than equip each phase and ground relay with its own fault detector.

Conclusion

There is reason to believe that distance ground relaying deserves much wider acceptance than it has received so far, since operating experience has demonstrated that distance ground relays give results which are in every way as satisfactory as those obtained with distance phase relays, the use of which is well established.

It also is believed that if more engineers would study the possibilities of distance ground relays in combination with distance phase relays and compare them with the much more expensive and complex carrier pilot relaying, they would find that distance relaying would give adequate protection in many cases where otherwise more expensive relaying would seem desirable.

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High Voltage Ignitron Rectifiers and Inverters for Railroad Service

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Synopsis: Single-anode pumped-type mercury-arc-rectifier tubes have been established in the 300-, 600-, and 900-volt field in the past few years, and as a result they have become the most important type of equipment for large power conversion at these voltages. The ignitron was the major development which brought about the change from multianode to single anode tubes in this low voltage field.[1] Recently sealed ignitrons have been developed for very high voltages and lower currents.[2] This paper describes high voltage pumped ignitrons which have been made available for power conversion in the 3,000- to 4,000-volt range. The major application for mercury arc rectifiers in this voltage range is for railroad service. Therefore, the requirements for railroad loading have been given special consideration in the design of the high voltage ignitron. Both rectifier and inverter characteristics were desired in one unit so that a maximum of flexibility could be obtained. The characteristics of combined rectifier and inverter operation are described. Accepted fundamental rectifier and inverter relationships are presented, and curves are included to aid in determining operating characteristics. The design and tests which led to the successful development of the high voltage pumped ignitron are discussed. Pumped ignitrons are compared with sealed ignitrons for high voltage applications so that the field of each can be better defined.

THE DEVELOPMENT of the 3,000-volt ignitron rectifier was begun after the characteristics of single anode tube and multianode mercury arc rectifiers were compared. In low voltage applications, the lower arc drop of single anode tube rectifiers results in an appreciable improvement in the efficiency and is the determining factor in the selection of the type of rectifier. In the high voltage field, the arc drop is a smaller percentage of the direct voltage, and the lower arc drop improves the efficiency only slightly. However, single anode tube rectifiers have a number of minor advantages that are important when considered together.

The maintenance of single anode tube rectifiers is simplified because they consist of a number of similar tubes. This fact also makes it possible to operate at reduced capacity with some of the tubes removed. Since large numbers of tubes are required, mass production methods can be used in the manufacture.

Ignitron rectifiers have the following advantages over continuously excited mercury arc rectifiers:

1. Cathode spot excitation independent of load.
2. Angle of retard unaffected by load current or water temperature.
3. No moving parts in excitation circuit.
4. No cathode insulation required.

Single anode tube rectifiers have the disadvantage that their excitation circuits are somewhat more complicated. However, since the excitation circuits are almost independent of the kw rating, it is found that this disadvantage is outweighed by the advantages for the higher ratings.

Ratings

Typical ratings required for rectifier units for railroad loading are 2,000 kw and 3,000 kw. These ratings were used as a basis for the design of the 3,000-volt ignitrons.

Railroad loading usually requires very heavy overload ratings. Ignitron rectifiers are especially suited for suddenly applied overloads, because they show no tendency to cause arc surges at any reasonable water temperature and because their arc-back performance is less dependent on previous loading.

The proposed AIEE heavy duty rating,[3] which is used in the testing of high voltage rectifiers for railroad service, is the following:

100 per cent rated load continuously

150 per cent current for two hours, following 100 per cent load

300 per cent current for five minutes, following 100 per cent load

With low voltage high current ignitrons, it is found that the inverter rating of a given design is lower than the rectifier rating. In high voltage rectifier operation, the required design results in improved inverter performance.

Regeneration

One of the important requirements for a railroad substation in mountainous country is regeneration. This requirement can be accomplished by mercury arc rectifiers in a number of different ways:

1. A given unit may be operated first as a rectifier when power is required and then as an inverter when regeneration is required. A reversing switch to change the polarity of the power circuit and means for rapidly changing the ignition angles must be provided. This scheme has the advantage that duplicate equipment is not required, but has the disadvantage that a time delay is necessary to change from rectification to inversion and that the reversing switch has considerable duty.

2. Two or more units may be operated on the same bus, with one permanently connected with reversed polarity and excited as an inverter and the others being operated as rectifiers. This method provides a smooth changeover between rectification and inversion, but has the disadvantage that duplicate equipment is required.

3. Where heavy overloads and heavy regenerative loads are expected, a combination of the first two methods can be used. Both units can carry the overloads, and as the load drops, one unit can be switched over to inverter operation to carry normal regenerative loads. If a heavy regenerative load is required, both units can be changed to operate as inverters. This type of operation makes possible a smooth transition between rectification and inversion.

Figure 1 gives a typical rectifier and inverter circuit illustrating the second method, which is considered to be the most important. Load and regenerative voltage characteristics are shown in Figure 4 for a system having identical transformers for rectifier and inverter units. These regulation lines give the direct voltages for various load currents and for different ignition angles α of rectifier and β of inverter. The slope of these lines is determined mainly by the transformer reactance, and the values at zero current are proportional to the cosines of the ignition angles.

Circuit Characteristics

General Theory

In normal operation the action of any type of rectifier is to allow energy flow in only one direction, because a reverse current cannot be conducted through the rectifier elements. However, the flow of

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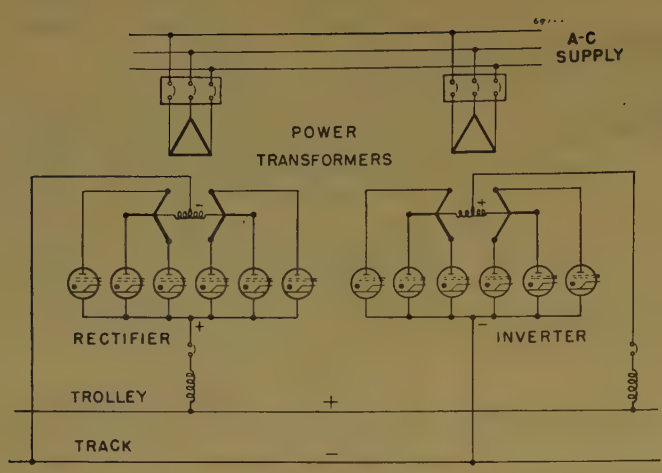


Figure 1. Delta 6-phase double-Y rectifier and inverter circuit for railroad substation

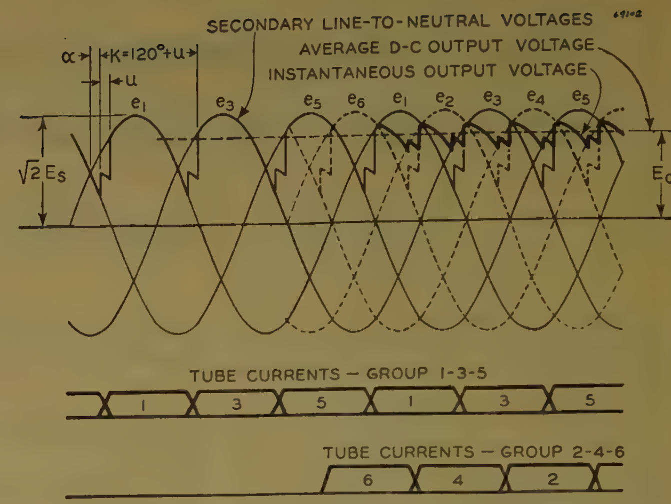


Figure 2. Operation of 6-phase double-Y rectifier

energy in any circuit element can be reversed by changing either the direction of current through the element or the polarity of the voltage across the element. In a converter in which the conducting period of the elements can be controlled, the polarity of the direct voltage can be reversed by making effective the opposite polarity of the alternating voltage. This action is the principle of operation of the inverter, which permits power to be transmitted from the d-c circuit to the a-c circuit.

With railroad service, the need for regeneration is indicated by a rise in the direct voltage. The effect of this rise on rotating equipment would be to reverse the direction of current flow, but with the rectifier the elements support the additional voltage and there can be no reverse current. A second converter with a reversed connection may be controlled so that the ignitrons can conduct during periods in which the alternating voltages of their transformer windings are opposed to the flow of current. With this type of operation, energy can be transmitted from the a-c circuit to the d-c circuit by the rectifier and from the d-c circuit to the a-c circuit by the inverter.

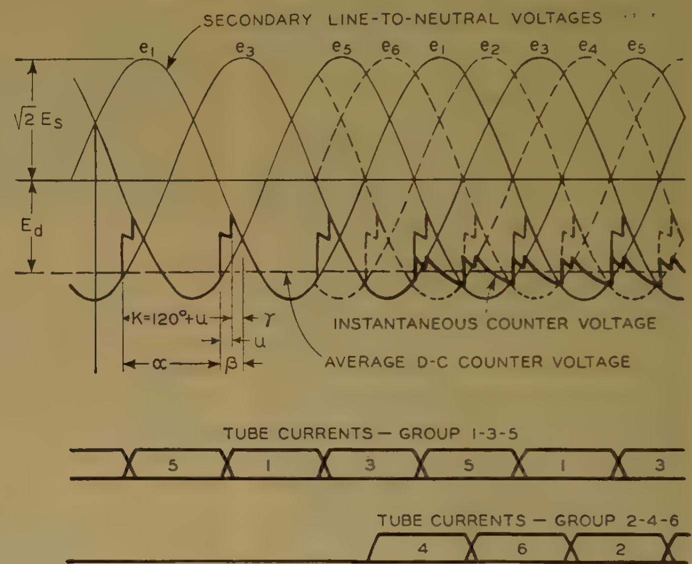


Figure 3. Operation of 6-phase double-Y inverter

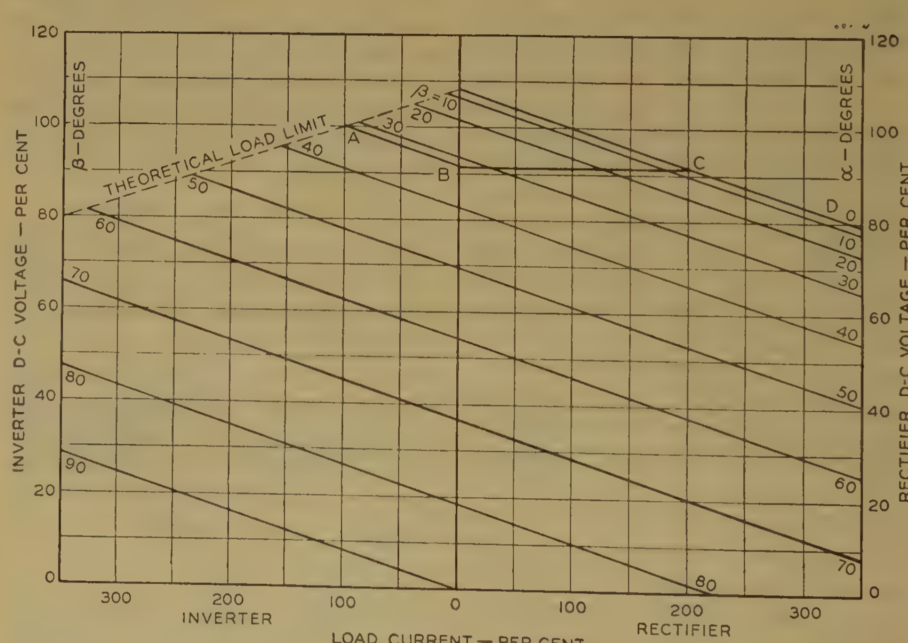


Figure 4. Regulation curves of rectifier and inverter, eight per cent regulation and same transformer secondary voltages

Rectifier Operation

In the simplified rectifier and inverter circuit shown in Figure 1, the rectifier operates as a normal delta 6-phase double-Y rectifier, since operation is not changed greatly by the addition of the inverter. Two separate transformers may be used as shown, or one transformer with a single primary may be used. The rectifier output voltage and current wave forms are shown in Figure 2.

The average value of the d-c output voltage of a rectifier at no load with ignition delay is*

E_{dc} = theoretical average value − average arc drop

$$E_{dc}=E_{do}\cos\alpha-E_a \qquad (1)$$

where

$$E_{do}=\sqrt{2}E_s\frac{p}{\pi}\sin\frac{\pi}{p}$$

The average value of the d-c output voltage of a rectifier under load is

E_d = theoretical average value − commutating reactance drop − resistance loss drop − average arc drop

$$E_d=E_{do}\cos\alpha-E_x-E_r-E_a$$
$$=\sqrt{2}E_s\frac{p}{\pi}\sin\frac{\pi}{p}\cos\alpha-\frac{p}{2\pi}I_cX_c-\frac{P_r}{I_d}-E_a \qquad (2)$$

or, using other quantities,

$$E_d=E_{do}\left[\cos\alpha-\frac{\cos\alpha-\cos(\alpha+u)}{2}\right]-\frac{P_r}{I_d}-E_a \qquad (3)$$

For a delta 6-phase double-Y circuit, $p=3$ and $I_c=I_d/2$. Therefore,

$$E_d=1.17E_s\cos\alpha-0.239I_dX_c-\frac{P_r}{I_d}-E_a \qquad (4)$$

For a delta 6-phase star circuit and a delta 6-phase fork circuit, $p=6$ and $I_c=I_d$. Therefore,

$$E_d=1.35\,E_s\cos\alpha-0.955I_dX_c-\frac{P_r}{I_d}-E_a \qquad (5)$$

Inverter Operation

Figure 3 shows the operation of a 6-phase double-Y inverter. Each anode conducts during the negative portion of the corresponding alternating voltage wave, in order to provide a d-c counter voltage opposing the flow of current.

The average value of the d-c counter voltage of an inverter at no load is

$$E'_{dc}=E'_{do}\cos\alpha'-E'_a$$

or

$$=-[E'_{do}\cos\beta'+E'_a] \qquad (6)$$

where

$$E'_{do}=\sqrt{2}\,E'_s\frac{p'}{\pi}\sin\frac{\pi}{p'}$$

The average value of the d-c counter voltage of an inverter under load is

$$E'_d=E'_{do}\cos\alpha'-E'_x-E'_r-E'_a$$
$$=\sqrt{2}\,E'_s\frac{p'}{\pi}\sin\frac{\pi}{p'}\cos\alpha'-\frac{p'}{2\pi}I'_cX'_c-\frac{P'_r}{I'_d}-E'_a \qquad (7)$$

or, using other quantities,

$$E'_d=E'_{do}\left[\cos\alpha'-\frac{\cos\alpha'-\cos(\alpha'+u')}{2}\right]-\frac{P'_r}{I'_c}-E'_a$$
$$=-\left\{E'_{do}\left[\cos\beta'+\frac{\cos(\beta'-u')-\cos\beta'}{2}\right]+\frac{P'_r}{I'_c}+E'_a\right\} \qquad (8)$$

For a delta 6-phase double-Y circuit,

$$E'_d=-\left[1.17\,E'_s\cos\beta'+0.239\,I'_dX'_c+\frac{P'_r}{I'_d}+E'_a\right] \qquad (9)$$

For a delta 6-phase star circuit and a delta 6-phase fork circuit,

$$E'_d=-\left[1.35\,E'_s\cos\beta'+0.955I'_dX'_c+\frac{P'_r}{I'_d}+E'_a\right] \qquad (10)$$

Combined Characteristics

Under normal operating condition of a rectifier and an inverter, as shown in Figure 1, the ignition angles must be such as to have the rectifier average d-c output voltage less than or equal to the inverter average d-c counter voltage, so as to prevent circulating currents. In the 6-phase double-Y circuit, there is a light load voltage rise in the rectifier and a light load voltage dip in the inverter. These changes occur when the load current becomes less than that required for magnetizing the interphase transformer, resulting in a transition to 6-phase conduction periods. These light load voltage changes can be eliminated by adjustment of ignition angles so that a small current circulates between the rectifier and inverter when there is no load on the system.

Figure 4 shows the combined characteristics of a rectifier and inverter for various ignition angles, with eight per cent regulation neglecting losses.

The rectifier curves have dropping voltage characteristics and are to the right of the zero current axis, and the inverter curves have rising characteristics and are to the left. The inverter curves terminate at the theoretical load limit line, where the commutation angles are exactly equal to the advance angles. In actual operation it is necessary to operate below the theoretical load limit so as to provide margin angles greater than tube deionization requirements.

When the transformer secondary voltages are the same for both rectifier and inverter, it is necessary to operate at high angles of delay in the rectifier to have

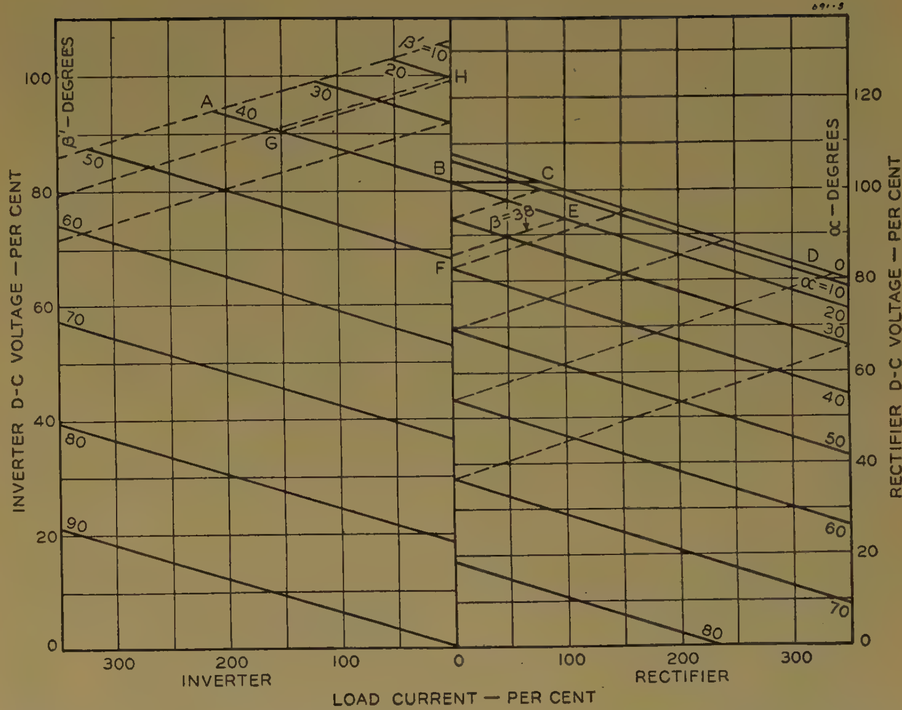


Figure 5. Regulation curves of rectifier with eight per cent regulation, and inverter with six per cent regulation and higher secondary voltage

* Definitions of the symbols are given at the end of this paper. Unprimed symbols are used for rectifier quantities and primed symbols are used for inverter quantities.

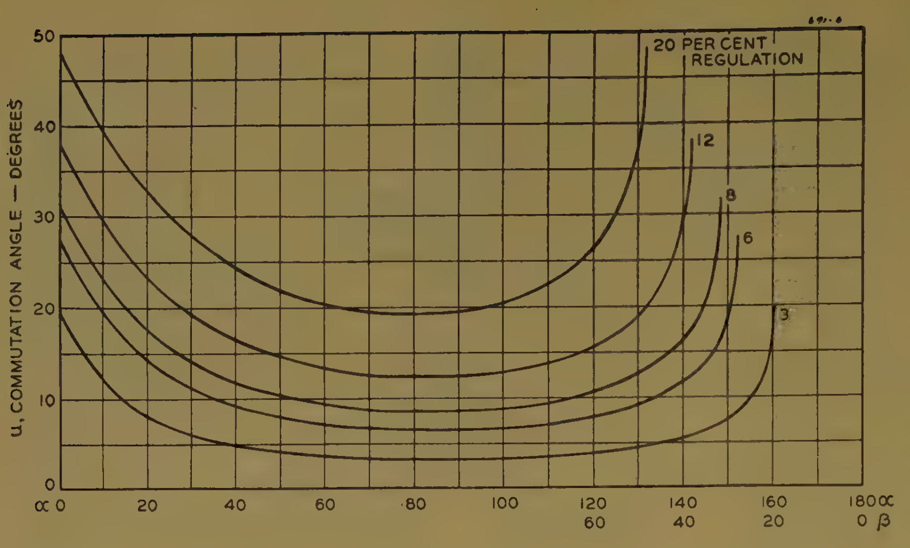


Figure 6. Commutation angles of a p-phase rectifier or inverter, neglecting losses

$$\text{Per cent regulation} = \frac{E_{do} - E_d}{E_d} 100, \; E_d \text{ at rated } I_d \text{ and } \alpha = 0$$

$$\frac{\cos\alpha - \cos(\alpha + u)}{2} = \frac{\text{regulation}}{1 + \text{regulation}}$$

α = angle of retard, degrees
β = angle of advance, degrees

sufficient advance angles in the inverter to carry the required regenerative loads. For the inverter to carry 100% current, the theoretical minimum advance angle β must be 32 degrees, shown along line *AB*. Actually less than 100 per cent current can be carried at this angle. The rectifier must operate at a minimum of 32 degrees delay at *B* to avoid circulating currents. As load is increased, the voltage may be kept constant by phase control from *B* to *C*.

Figure 5 shows a condition which is more practical. The rectifier is not required to operate at high angles of delay, by having a higher transformer secondary voltage on the inverter. The direct voltage of the rectifier is held constant between *B* and *C* by phase control as the load increases. With further increase in current, the voltage will drop along the α = 0 line from *C* to *D*.

Regenerative loads will be carried by the inverter with a d-c counter voltage given by the line *AB*, for β = 40 degrees. Any desired voltage characteristic can be obtained by automatically changing the firing of the inverter as the load changes, providing care is taken to operate at a sufficient amount below the theoretical load limit line.

A method of making reasonably accurate determinations of rectifier voltage, ignition angles, and commutation angles under various load conditions is by use of imaginary inverter regulation lines superimposed on the rectifier lines. These lines are shown dotted and of equal slope as the rectifier lines, but with rising characteristics. The advance angles β of these lines are equal to the delay angles α of rectifier lines starting from the same point on the zero current axis. From the symmetry of the rectifier and inverter voltage wave forms (Figures 2 and 3), the imaginary β values correspond to α plus u values of the rectifier.

The direct voltage of the rectifier at rated current and 20 degrees delay can be found by moving down on regulation line *BE* for α = 20 degrees from *B* to 100 per cent current at *E*, and reading 93 per cent rated voltage opposite *E*. The imaginary inverter line *EF* gives β = 38 degrees by interpolation between the 30- and 40-degree lines. The rectifier commutation angle u is equal to 38 minus 20, or 18 degrees.

Inverter conditions can be determined in a similar manner by using the dotted imaginary rectifier lines. For example, the inverter counter voltage for 150 per cent current at 40 degrees advance is given as 90 per cent at point *G*. The imaginary rectifier line *GH* gives α = 21 degrees as the imaginary rectifier ignition angle; this value corresponds to the inverter margin angle γ. The inverter commutation angle u then is β minus γ, or 40 minus 21 equals 19 degrees.

Choice of Circuit

Most of the 3,000-volt mercury arc rectifiers which have been applied up to the present time have used 6-phase star or fork transformer secondaries. The important reason for this use has been to avoid the light load voltage rise of the double-Y circuit. However, in stations in which there are both rectifiers and inverters, the light load rise can be eliminated as described previously. For a given rated direct current and for equal direct voltage regulation, the 6-phase double-Y circuit has the following advantages over the star or fork connections:

1. Peak anode currents are half.

2. Rms anode currents are lower by ratio of $\sqrt{2}$.

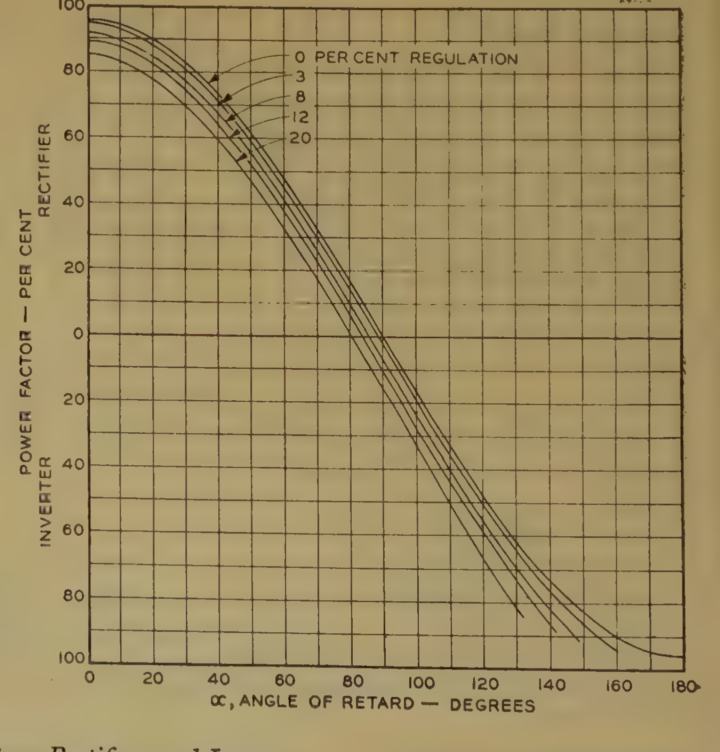


Figure 7. Power factor of 6-phase rectifier or inverter at rated current, neglecting exciting currents and losses

3. Transformer kilovolt-amperes is lower.

4. Final commutation rate of anode current is lower by the ratio of 2.

5. Short circuit and arc back currents are lower.

6. Low transformer commutating reactance is not required.

The 6-phase star or fork circuits have the following advantages:

1. Initial inverse voltage is lower by ratio of 2.

2. There is no light load voltage rise.

The circuits are the same in the following:

1. Average anode currents are equal.

2. Peak inverse voltages are equal.

3. Commutation angles are equal.

4. Power factors are equal.

Figure 6 shows the commutation angles of any type of rectifier circuit as a function of the angle of retard and the d-c regulation, neglecting losses. Figure 7 shows the theoretical power factor of the 6-phase double-Y, the star, and the fork circuits as a function of the angle of retard and voltage regulation.

Faults

The important type of electronic fault with rectifier operation is the arc-back, which is defined as the failure of the tube to hold the negative anode to cathode voltage, with the result that a current flows in the reverse direction because of the formation of a cathode spot on the anode. An arc-back on one anode produces a short circuit on both the d-c and a-c circuits and must be cleared by opening both the a-c and d-c breakers. With high voltage rectifiers, the tubes may be blocked by electronic arc suppression to prevent the flow of forward current in the unfaulted tubes. However, the d-c breaker must be opened to interrupt the flow of current from any equipment capable of sustaining a direct current through the faulted tubes.

If a rectifier is operating with voltage reduction by ignition delay, the positive anode to cathode voltage may cause breakdown of the tube before it is released. This type of fault is known as a forward-fire. If only one such fault occurs, it is self-correcting without opening any circuit breakers, but if all tubes continue to forward-fire, the output voltage will be increased and the unit may be overloaded. With ordinary water temperatures and the tight shielding used on high voltage rectifiers, faults of this type almost never occur.

Another general class of faults is the misfire, which is the failure of a tube to begin conducting at the proper time. An occasional misfire in a rectifier will not cause any serious disturbance, but a series of misfires will result in the lowering of the average output voltage.

In inverter circuits the voltages are such that an arc-back will not cause a tube to continuously carry current in the reverse direction. An inverter tube may have an arc-back caused by the negative voltage on the anode at the end of the conduction period, but the voltage on the tube reverses and stops the flow of the fault current. An arc-back usually will lead to the arc-through, which is the more serious fault of the inverter.

An arc-through is defined as the conduction of power circuit current in the forward direction through a tube during a scheduled nonconducting period of the cycle. An arc-through may be caused by one of the following:

1. Forward-fire, the failure of a tube to block the forward voltage on the anode. This tube fault may or may not produce a circuit fault.

2. Misfire, the failure of a tube to begin conducting current at the start of the scheduled conducting period, thus producing an arc-through on the preceding tube. It is a fault caused by either the excitation circuit or the tube and always produces a circuit fault for at least one cycle.

3. Recommutation, the failure of the load current to be commutated completely from one tube to another within the required angle, with the result that current is commutated back to the original tube. It is a fault caused by the power circuit.

If a rectifier and an inverter are operating on the same bus and an arc-back occurs on the rectifier, the inverter will

Figure 8. Single-grid high-voltage ignitron

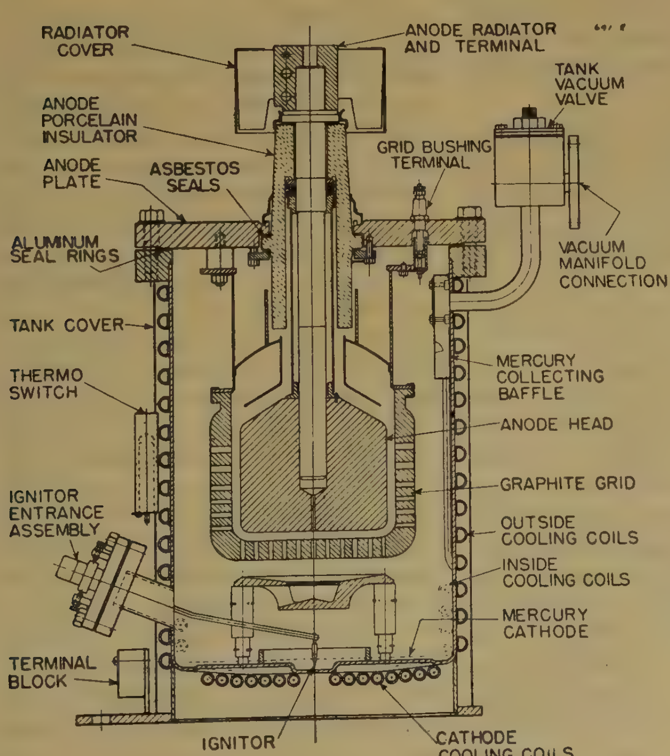
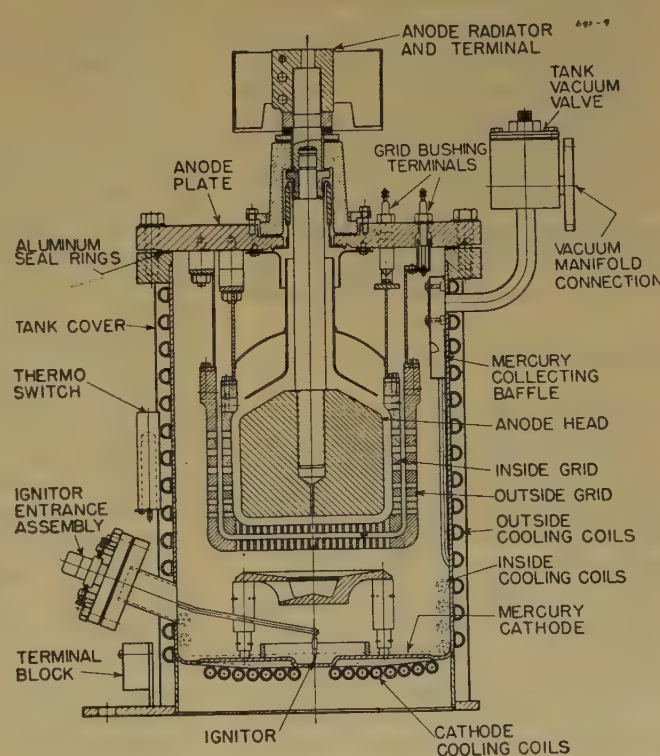


Figure 9. Double-grid high-voltage ignitron



not feed current into the short circuit. However, if an arc-through occurs on the inverter, the rectifier will supply a fault current to the inverter. The other tubes in the inverter, which do not have a fault, do not supply any current to the faulted tube or tubes. If the units are identical, the peak arc-back currents are about equal to the peak arc-through currents.

Ignitron Design

In the design of the 3,000-volt pumped ignitron, attention was given to obtaining a rectifier with a low arc-back rate and high overload capacity. The operating outlet water temperature was to be 55 degrees centigrade. The same design was to be used for either the rectifier or the inverter, and the shielding of the anode was to be such that only ignitor control would be required for either type of operation.

To accomplish the above objectives, two designs of ignitrons were made at about the same time. Both of these were manufactured and placed on test. The results were compared, and the best design was selected and given additional tests.

The first design of ignitron used a single grid with a high ion attenuation and with a high deionization influence around the anode. A porcelain anode bushing with a soldered seal was used on this design. Figure 8 shows the cross section of this single grid design.

The second design used two grids having a total deionizing effect about the same as that of the single grid in the first design. The grids were mounted separately and a circuit was connected to each through insulating bushings. The original design of the double grid ignitron had a porcelain bushing with a soldered seal, and the first tests were made with this type of bushing, but later an improved type of bushing was used. This new bushing and the final design of the double grid tube are shown in Figure 9.

Tests

Single Grid Design

Power Circuit. The power circuit used in the tests of the high voltage ignitrons is shown in Figure 10. The four d-c machines were used either as a load for rectifier tests or as a d-c supply for inverter tests. This arrangement of motor-generator sets provided a continuous rating of 2,000 amperes at 3,000 volts. In order to change from rectifier testing to inverter testing, it was necessary to reverse the d-c power circuit and to reverse the polarity of the holding coil in the high speed d-c breaker.

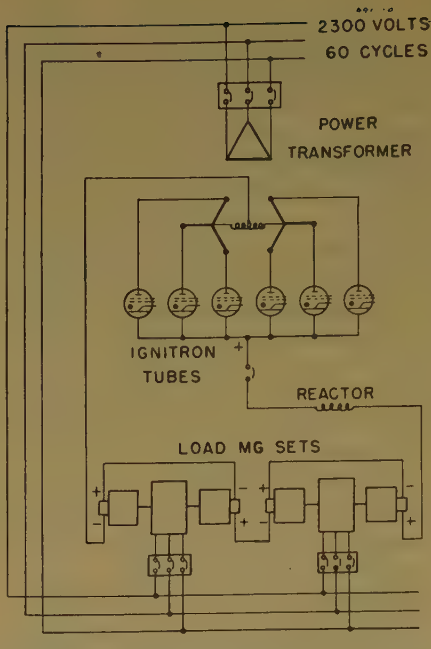


Figure 10. Circuit for laboratory tests of high voltage ignitrons

Excitation Circuits. At the start of the high voltage tests, standard firing and grid circuits were used. The grid circuit soon was found to be unsatisfactory for the 3,000-volt ignitron, as there was difficulty in the pickup of the main anode. At normal operating temperatures, the firing of the ignitors and the pickup of the grids were satisfactory, but the main anode had frequent misfires and required high minimum load current for pickup. This difficulty was corrected by increasing the grid circuit current by a considerable amount.

Power Circuit Faults. In the original 3,000-volt rectifier tests, it was found that a high arc-back rate was obtained. The faults of the single grid design and the original double grid design had about the same characteristics, and they appeared to be produced by the same causes.

With rectifier operation the arc-backs seemed to be the result of entirely different causes than are experienced in low voltage mercury arc rectifiers. The arc-backs were almost independent of load current and were found to occur even at no load. It was believed that this characteristic was caused by mercury condensed on the cold anodes. This theory was eliminated, however, when it was possible to make several tests in which the rectifier operated without arc-backs for several hours and then had a series of arc-backs in a few minutes. Operation at reduced voltage to warm the ignitrons, and then change to the high voltage did not improve the performance and indicated that mercury on the anodes was not the major cause of the arc-backs.

One of the important clues as to the cause of arc-backs was the phase angle in the inverse voltage period at which they occur.[7,8] The most accurate and complete data were obtained by the use of the memnoscope.[9] A memnoscope oscillogram of the anode to cathode voltage of three tubes is shown in Figure 11. The fault occurred near the peak inverse voltage, where the anode to cathode voltage dropped to zero. In most of the records taken of the high voltage ignitrons, the faults occurred near the peak of the inverse voltage. The breakdown characteristics of the ignitrons were investigated with no ionization in the tubes It was found that a new tube just degassed would break down at slightly below its rated inverse voltage. However, if the testing was continued with repeated breakdowns, the voltage held between the anode and cathode could be increased to many times the rated value.

One of the first phenomenon which was investigated was the effect of mercury dew on the surface of an insulator. It was thought that enough mercury could be condensed to reduce substantially the value of the creepage distance on the insulator. A test setup was made in which a voltage stress could be placed on an insulator with a controllable temperature in mercury vapor at pressures corresponding to those in an ignitron. This test indicated that it took a large amount of mercury condensed on the surface of an insulator to reduce the creepage breakdown voltage even a slight amount. These tests were definite enough to indicate that this phenomenon was not the cause of the breakdowns in the ignitrons.

Another possibility that was investigated was the effect of mercury drops at the junction of the insulator and the conductor. It was believed that there might be a condition similar to that existing between an ignitor and the mercury pool, which would make it possible to start the cathode spot of the arc with potential gradients that appeared to be low. An investigation of the conditions at the junction of the anode stem and the insulator, and of models with similar arrangements, indicated that it was very probable that this junction was a source of the cathode spots which were producing the arc-backs. The junction was placed close to the anode plate, and it appeared that there was enough capacitance between the mercury drops and

the anode plate to produce cathode spots at the junction. Another, but less likely, explanation was that high resistance ignitor action took place at the junction, even though most of the insulator did not have mercury on it.

Since condensed mercury was the cause of the arc-backs, bushing heaters were used so that the junctions were held at a temperature at which mercury would not condense. An interesting fact was observed that at higher loads, higher bushing temperatures were required to prevent the start of a series of arc-backs. Heavy overloads frequently were found to produce arc-backs which occurred 10 or 15 minutes after the overload was completed. These arc-backs also were corrected by operating with a sufficiently high bushing temperature.

In the investigation of the probable causes of arc-backs, a high voltage ignitron was built with several windows in the tube. The conditions in the tube during the various portions of the cycle were investigated by means of a stroboscope. One observation was that large numbers of mercury drops were moving in the grid-cathode space at high speed. It was believed that if one of these mercury drops hit the hot anode during the negative portion of the cycle, there would be a high probability for an arc-back. By using a special design of grids and baffles it was possible to shield the anode from the drops. However, it is still possible that this factor may be one of the important causes of arc-backs in all mercury arc rectifiers.

Double Grid Design

Power Circuit. The power circuit used in testing the double grid design was the same as that used for the single grid design and is shown in Figure 10. For some of the rectifier tests and for most of the inverter tests, the d-c inductance was increased so that there was less ripple in the direct current.

Excitation Circuit. The use of a double grid tube greatly increased the number of types of excitation circuits which could be used. However, it was desirable to keep the excitation circuit as simple as possible. With the double grid ignitrons, there was little trouble with pickup, either of the inside grid or the anode, because the ion attenuation of each grid was not unusually high. For rectifier operation, it was found that the most desirable grid circuit was with sine wave excitation at a low voltage, and with angles of retard determined by the ignitors.

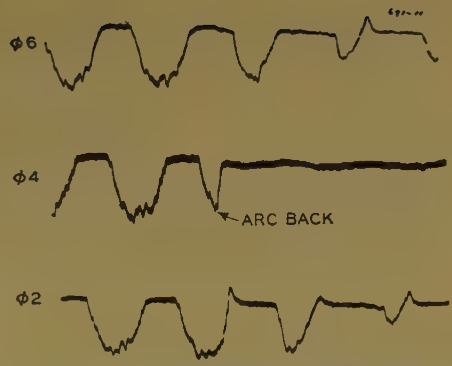


Figure 11. Oscillograms of anode-cathode voltages, showing arc-back in phase 4

For inverter operation, the problem of excitation was found to be more difficult. For inverter operation with normal overloads, the same low voltage grid circuit and ignitor control could be used as in rectifier operation. However, if unusually high overloads were required on the inverter, it was necessary to use a grid circuit in which the outer grid released the inner grid when the ignitor was fired.

Rectifier Tests. The rectifier tests with double grids and porcelain-enclosed glass kovar sealed bushings were considered very satisfactory. The new bushing was found not to be critical to the collection of mercury dew, as its permissible operating temperature was far above that at which mercury would condense in the tube. This condition made it possible to eliminate completely the bushing problem, since a high temperature could be maintained by the use of anode and bushing heaters.

The use of a double grid greatly increased the load current which could be carried with a permissible arc-back rate. By operating the grids at low voltage, forward-fires practically were eliminated, even when operating at high angles of ignition delay. The rating obtained for the rectifier was higher than the required rating.

Inverter Tests. The double grid ignitrons operated as inverters with ignitor control, making it possible to have a simple excitation circuit. The limiting rating of the inverter was less than that of the rectifier.

Most of the faults which occurred in the inverter were arc-throughs caused by forward-fires. The memnoscope oscillograms indicated that the loss of control usually occurred just after the anode became positive. Rarely, a recommutation fault was observed when there was a disturbance in the a-c line voltage caused by some other equipment on the shop bus.

Sealed Ignitron Tubes

The possibility of using sealed ignitron tubes for the 3,000-volt class should be considered for some railroad applications. Sealed ignitrons have been developed for voltages up to about 7,000 volts d-c in single way circuits, and economical designs could be obtained for 3,000 volts.

There should be no fundamental difference in the performance of pumped and sealed ignitrons. Up to the limit of size for which experience has been obtained in sealed tube construction, any electrical design used in pumped tubes also can be used in sealed tubes with an equal rating.

The decision as to whether pumped or sealed ignitrons should be used in a given application should be based on the relation between the required continuous and overload ratings and the available tube ratings. Economic considerations may be the determining factor for some current capacities.

Conclusions

High voltage pumped ignitrons of an improved design have been built and tested for the requirements of 3,000- to 4,000-volt railroad service. All of the circuit characteristics, which are available with grid controlled rectifiers and inverters, are obtained by the use of ignitor control. Rectifiers and inverters can be used in one substation to provide the desired characteristics for either load or regeneration.

The high voltage ignitrons are designed for heavy overloads of the type required for railroad service. A new design of porcelain-enclosed glass kovar sealed anode bushing, which has a temperature limit far above that obtained in normal operation, has been developed.

The selection of either pumped or sealed construction for high voltage ignitrons must be based on the rating, economic factors, and special requirements for each application.

Nomenclature: Rectifier Symbols

a_m = coefficient of sine term of Fourier expansion for the mth harmonic (crest value)

b_m = coefficient of cosine term of Fourier expansion for the mth harmonic (crest value)

c_m = coefficient of resultant of Fourier expansion for the mth harmonic (crest value)

E_a = average direct voltage drop caused by loss in the converter elements (average arc drop)

E_d = average direct voltage under load

E_{dc} = average direct voltage, no load, with phase control and with arc drop

E_{d0} = theoretical direct voltage (average direct voltage at no load assuming no overlap, no phase control, and zero arc drop)

E_{ti} = initial inverse anode to cathode voltage

E_L = a-c system line-to-line voltage in rms volts

E_N = a-c system line to neutral voltage in rms volts

E_{pf} = peak forward anode to cathode voltage

E_{pi} = peak inverse anode to cathode voltage

E_r = average direct voltage drop caused by resistance losses

E_s = transformer direct winding line to neutral voltage in rms volts

E_x = average direct voltage drop caused by commutating reactance

ΔE = total average direct voltage drop because of regulation

f = frequency of a-c power system

$F_x = \frac{I_c X_c}{E_s} = 2\sqrt{2} \sin \frac{\pi}{p} \frac{R_{gx}}{1+R_{gx}}$ = reactance factor

I = a-c line current (crest value)

I_a = average anode current

I_c = direct current commutated between two rectifying elements in a commutating group

I_d = average d-c load current

I_e = transformer exciting current in rms amperes

$I_H = \sqrt{I_5^2 + I_7^2 + I_{11}^2}$ = equivalent totalized harmonic component of I_L

I_L = a-c line current in rms amperes

I_1 = fundamental component of I_L

I_{1P} = watt component of I_1

I_{1Q} = reactive component of I_1

I_m = harmonic component of I_L, of the order m

I_p = transformer a-c winding current in rms amperes

I_s = transformer d-c winding current in rms amperes

L_d = inductance in the d-c reactor in henries

m = order of harmonic

p = number of phases in a simple rectifier

P_r = total copper resistance losses in watts

q = total number of rectifier phases

$\underline{R}_g$ = per-unit direct voltage regulation based on full load voltage with no phase control

$\underline{R}_{gx}$ = per-unit direct voltage regulation caused by reactance based on full load voltage, with no phase control and assuming no losses

R_p = ohmic resistance of the a-c winding

R_s = ohmic resistance of the d-c winding

u = commutation angle (angle of overlap)

X_c = commutating reactance line to neutral in ohms

X_L = ohms reactance of supply line (per line)

$\underline{X}_L$ = per-unit reactance of supply line (expressed on basis of rated volt-amperes at the line terminals of the transformer a-c winding)

$\underline{X}_T$ = per-unit reactance of transformer (expressed on basis of rated volt-amperes at the line terminals of the a-c windings)

α = angle of retard (delay angle, ignition angle; the angle by which the start of conduction lags the point where the incoming and outgoing rectifying elements have equal positive voltages)

$\cos \alpha$ = phase control ratio

β = angle of advance (the angle by which the start of conduction leads the point at which the incoming and outgoing rectifying elements have equal negative voltages)

γ = margin angle (the angle by which the end of commutation leads the point where the incoming and outgoing rectifying elements have equal negative voltages)

$\cos \phi$ = displacement component of power factor

$\cos \delta$ = distortion component of power factor

κ = conduction angle for one rectifying element

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Modern Double-Reduction Traction Motor

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Synopsis: Incorporated in the modern double-reduction motor and gear unit are many new design features which have made it an outstanding motor for the past five years. Oil lubrication is provided for all gearing and bearings in the oil-tight double-reduction gear unit. Special design constants, which have been developed as a result of many years of experience with high-speed motors, are used. Self-ventilation is successful on a high-speed motor and simplifies locomotive design. Specially insulated field coils provide for good heat transfer from the coils to steel and air. A small motor with a high-speed armature results in low copper loss and high efficiency at high tractive effort. A steep speed curve with a large amount of field shunting makes it possible to use all motors connected permanently in parallel. The motor is small, compact, and lightweight, which has made possible its use on a wide range of locomotives. Here is a small motor doing a big job.

As EARLY as 1931, some thought was being given to the use of lightweight high-speed motors on gas-electric rail cars. The idea was to adapt motors which were designed for gas-electric drive on busses, for use in rail transportation. The first rail applications used a motor mounted either on the truck or body of the car. The motors were not designed sufficiently rugged to withstand the shock of operation when axle-mounted. On these applications automotive-type double-reduction or worm gearing was used, with a propeller shaft and double universal joints connecting the gearing and the motor. In 1938 a similar scheme was tried on lightweight industrial Diesel-electric switching locomotives. The 23-ton and 43-ton locomotives were the first on which this type of drive was used. These locomotives were equipped with a high-speed motor and double-reduction gearing.

The GE-733 is a lightweight double-reduction axle-mounted motor designed especially for Diesel-electric switching locomotives. The first of these was designed and built in 1940, and since that time a large number of these motors have been built and are in service all over the world. This machine was designed to cover a wide range of applications and therefore was built with many new electrical and mechanical features which have made it capable of doing the job well for the past five years. Figure 1 shows the motor complete with double-reduction gear unit.

More Tractive Effort

The primary objective in traction motor design is pulling power or tractive effort. It is the constant aim of the designer to secure more continuous tractive effort per pound of motor weight in each new design. If this can be done, the resulting small motor can be used on small-diameter wheels and on narrow gauge locomotives. Such a motor can be adapted to a wide range of locomotive sizes, simplifying their design. For example, weight saved in the traction motor may be used in the Diesel engine to make a more powerful locomotive without increasing the weight.

Figure 2 shows continuous tractive effort per pound of motor weight for various motors. Column *A* shows single-reduction motors which were being built in the period 1925 to 1930. Column *B* shows single-reduction motors which are being built today. Column *C* shows an early design of double-reduction motor which was built in 1936. Column *D* shows the GE-733 lightweight double-reduction motor as it now is being built and used on a wide variety of locomotives. The continuous tractive effort per pound of motor weight for the various motors shown on the chart in Figure 2 is based on the maximum reduction gearing which will give clearance under the gear case, as specified by the Interstate Commerce Commission, with the minimum wheels for which the motor is designed. This chart shows that steady progress has been made in the more efficient use of materials in both single-reduction and double-reduction motors. The continuous tractive effort per pound of the lightweight double-reduction motor is over $3^1/_8$ times as much as that of the 20-year old single-reduction motors and 43 per cent greater than that of the earlier design of double-reduction motor. This more efficient use of material is accomplished by using a high-speed lightweight motor and high torque multiplication in the double-reduction gear unit.

Figure 3 shows a graph of the continuous tractive effort and motor weight for the same groups of motors that are shown in Figure 2. For instance, the lightweight double-reduction motor weighs only one-third as much as a comparable present-day single-reduction motor and has 70 per cent as much continuous tractive effort rating. To judge a motor by this chart, place a straightedge along the abscissa and rotate it counterclockwise toward the ordinate, using zero as the center of rotation. The better motors, so far as continuous tractive effort is concerned, are nearer the ordinate.

How Much Horsepower?

A motor cannot be judged solely by its continuous tractive effort. Consideration also must be given to locomotive speed. If it were not necessary to consider locomotive speed as well as continuous tractive effort, triple-reduction or quadruple-reduction gearing could be used. This would give much higher

Figure 1. GE-733 d-c traction motor with double-reduction gear unit

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torque multiplication and therefore a corresponding increase in tractive effort, since tractive effort is motor torque translated to the rim of the locomotive wheel. The data given in Figure 4 take into consideration locomotive speed as well as tractive effort. These factors have been combined to obtain a measure of horsepower. This has been done by the approximate formula

$$\text{Horsepower} = \frac{\text{continuous tractive effort}}{375} \times \frac{\text{maximum miles per hour}}{3}$$

The reason for the "3" in the formula is that continuous tractive effort and maximum speed never occur at the same time in a Diesel-electric locomotive. In general, the continuous tractive-effort rating comes somewhere near one-third of the maximum speed. Any motor may be used with various capacity engines, but this formula makes possible a fair comparison between motors. The chart is shown in horsepower per thousand pounds of motor weight, and is shown for both single-reduction and double-reduction motors. The nomenclature of the columns in Figure 4 is the same as that used in Figure 2. As indicated by the height of the columns, much progress has been made in the past 20 years in the design of single-reduction motors. Also, the modern double-reduction motor shows approximately 20 per cent more horsepower per thousand pounds of motor weight than the earlier double-reduction motor, and 67 per cent as much as the modern single-reduction motor. This, together with the data shown in Figures 2 and 3, indicates that there is a place for double-reduction motors in light-traction road service, as well as in industrial switching service, where maximum power and operating speeds are not too high. However, on the heavier high-speed road locomotives, the horsepower to be handled is so great that large-diameter armatures are necessary for commutation. These large diameters make it necessary to keep the revolutions per minute low on account of the centrifugal forces involved. With low revolutions per minute, there is no advantage in having double-reduction gearing, because all of the reduction needed can be built into one set of gearing.

New Double-Reduction Gear Unit

The specially designed 1-piece cast steel gear case which forms a part of the motor and gear unit is shown in Figure 5. This is a view of the axle and

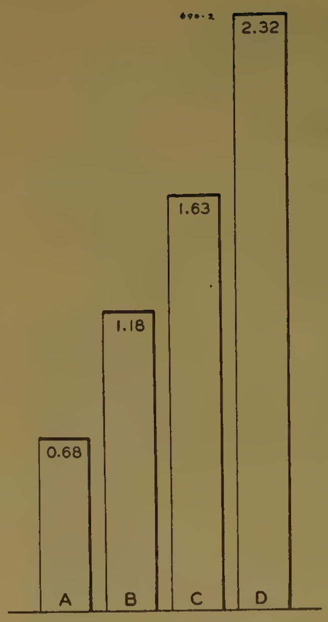
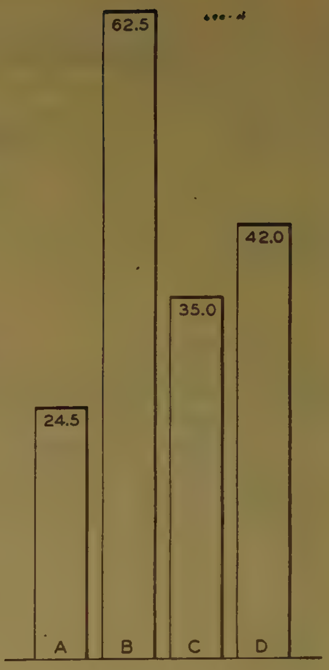


Figure 2 (left). Continuous tractive effort per pound of motor weight

Figure 4 (right). Horsepower per thousand pounds of motor weight

$$\text{Horsepower} = \frac{\text{continuous tractive effort}}{375} \times \frac{\text{max mph}}{3}$$



motor side and shows axle preparation, gear cover bolting flange, intermediate gear shaft bearing opening, bolt circle for mounting the motor, and air outlets for the motor ventilating fan. The gear cover is a separate piece and is cast of malleable iron. It is bolted to the gear case along the bolting flange. Figure 6 shows a photograph of the double-reduction gear unit with gear cover and axle caps removed. This illustration shows the gear unit completely assembled with all gearing and bearings in place. All bearings and pinions are arranged so that they can be disassembled without removing the motor and gear unit from the locomotive. Also, the complete motor can be removed and still leave the gear unit mounted in the locomotive. Thus, by using a high-speed motor, the size of the motor has been reduced to such an extent that it has become a part of the gear unit rather than the gear unit being a part of the motor, as on conventional

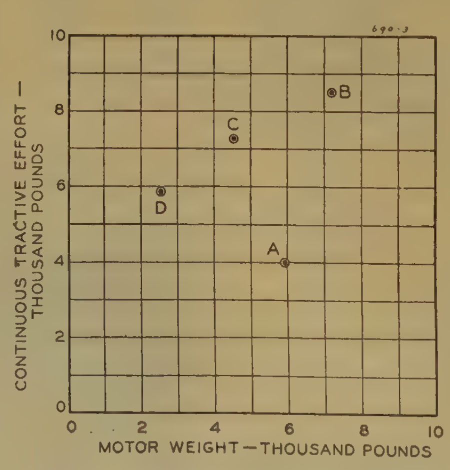


Figure 3. Continuous tractive effort and motor weight

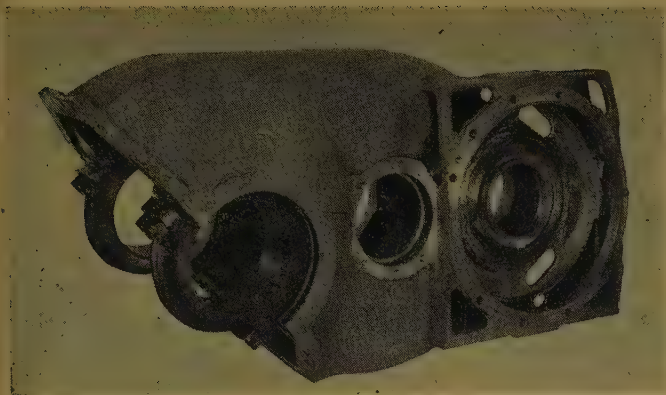
single-reduction machines. Figure 1 shows the motor and gear unit nose support which is located so that, for either direction of motion of the locomotive, there will be no tilting to throw the gear mesh out of alignment or place a twisting load on the axle bearings under the heavy tractive efforts encountered in service. Hardened spur gearing operating in an oil-tight gear case has a long life. Both the high-speed and low-speed pinions are straddle-mounted to maintain alignment.

The flood-lubricating system in the gear unit uses oil of about the viscosity of SAE 60 for normal operating temperatures, and lighter oil for lower temperatures. This lubricates all bearings in the gear unit, the pinion end motor bearing, the sleeve axle bearings, and thrust surfaces, as well as all of the gearing.

Mechanics of a High-Speed Motor

The motor is self-ventilated, which eliminates motor blowers and thus simplifies the locomotive design. This is found to be practical in a motor which is used with double-reduction gearing, since even in slow-speed switching service motor speeds are high enough to make the motor fan effective. The fan, which is mounted on the pinion end of the armature shaft, provides for multiple ventilation of the machine. One air path is over the commutator, through the space between the field coils and over the armature surface; the other is under the commutator and through longitudinal ducts in the core and armature heads.

A tight cylindrical smooth commutator is essential to the successful operation of



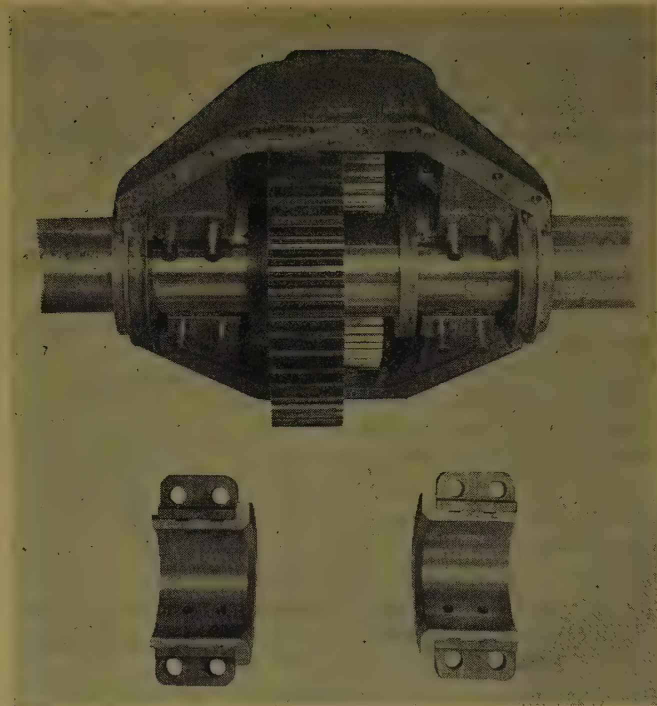


Figure 5 (above). Gear case for double-reduction gear unit

Figure 6 (right). Double-reduction gear unit with gear case and axle caps removed

any high-speed d-c machine. This is obtained by using commutator design constants, which have been developed over a period of years of experience with high-speed motors. Commutators are seasoned after the armature is completely assembled. This seasoning consists of subjecting the commutator to centrifugal forces and thermal stresses in excess of those which are encountered in service and tightening, and grinding the commutator until it operates smoothly, both hot and cold.

The armature is designed to withstand the stresses of high-speed operation that go with double-reduction gearing. The coils are held in the core slots by wedges, and the end windings are held with binding wire. The completed armature is dynamically balanced to close limits so that it operates smoothly at all speeds. It is also equipped with anti-friction bearings on both ends. The commutator end bearing housing is designed so that the armature can be removed without opening the bearing housing and exposing the bearing to dirt. This bearing also is arranged with all-metal labyrinth seals to keep grease in and to keep dirt out. The bearing on the pinion end of the motor armature is mounted outside of the motor pinion to form a straddle-mounting for the pinion. The bearing on this end is housed in the gear case and lubricated with the same oil as the gearing. The commutator end bearing is grease-lubricated.

The frame of the motor is bolted to the gear case so that the two become an integral unit and are mounted as such. Figure 1 shows a photograph of motor and gear unit.

Electrical Features

High-temperature insulating materials, such as asbestos cloth, asbestos paper, fiber-glass, and mica, are used in the modern double-reduction motor. With the aid of modern synthetic high-temperature varnishes, asbestos and mica, as well as fiber-glass and mica, have been combined to form tapes and wrappers. These materials have made it possible to use thinner insulation of a better quality and get a better-insulated machine. Thinner insulation leaves more space for copper and iron and at the same time improves the heat transfer coefficient through the insulation. Another advantage of these materials is their ability to stand high temperatures and at the same time have long life. High temperatures make it possible to get more tractive effort rating per pound of material. Inherent long life of insulating materials keeps maintenance costs low.

Proper proportioning of copper and iron in the magnetic circuit has made it possible to reduce materially the size and weight of the field structure. The field coils have been reduced in size by the use of new insulating materials and methods. Both exciting and commutating coils are edgewise wound and formed to fit the poles. They are insulated between turns with asbestos paper. A sufficient number of turns on the armature and frame sides of the coils are taped with mica and fiber-glass to provide the required insulation. Figure 7 illustrates an exciting and a commutating field coil with this type of insulation. The commutating coil is shown before the asbestos insulation has been inserted between the turns and before the coil has been dipped in high-temperature insulating varnish. The exciting field coil is shown completely insulated and ready to be mounted on the pole piece. Mica insulating collars around the pole pieces protect the field turns that are not taped. The coils are mounted on the poles and given a number of dips and bakes in high-temperature synthetic insulating varnish; thus the pole and coil become a complete unit and are assembled in the motor as such. With this type of coil, the heat transfer coefficients, both to air and iron, are much higher than for a conventional coil. After the poles and coils are assembled in the frame, the complete assembly is dipped in modern synthetic varnish and baked to fill completely and insulate all connections and joints so that they are sealed and protected from water, oil, and other foreign materials that may enter the motor.

Figure 7. Exciting and commutating field coils for modern double-reduction motor

The design constants of the motor have been made to produce a steep speed curve which, with the maximum amount of field shunting, makes it possible to use the motors connected permanently in parallel. This motor connection reduces wheel slippage and gives the locomotive more pulling power. Figure 8 shows characteristic curves of the GE-733 motor on 250 volts and for two field strengths. The wide change of flux in the motor, between maximum tractive effort and maximum permissible speed, together with the engine generator characteristics, makes a locomotive with a wide range of full utilization of the available engine horsepower.

High Efficiency

The power of a Diesel-electric locomotive is limited to that which the Diesel engine can deliver. For this reason, it is necessary that the transmission efficiency be kept as high as is compatible with reasonable design. This is especially true on switching locomotives where small engines are used and where heavy loads are handled at low speeds and high

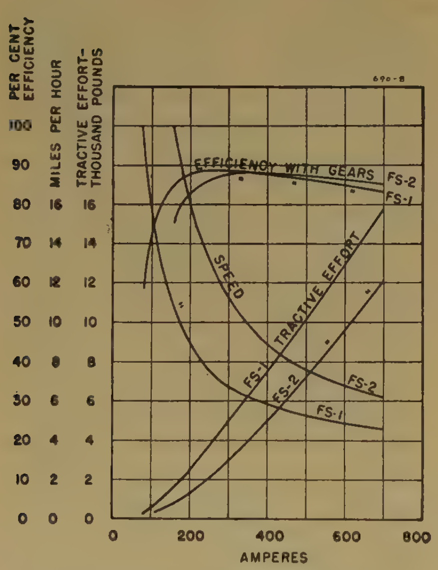


Figure 8. Characteristics of GE-733 motor on 250 volts and maximum reduction

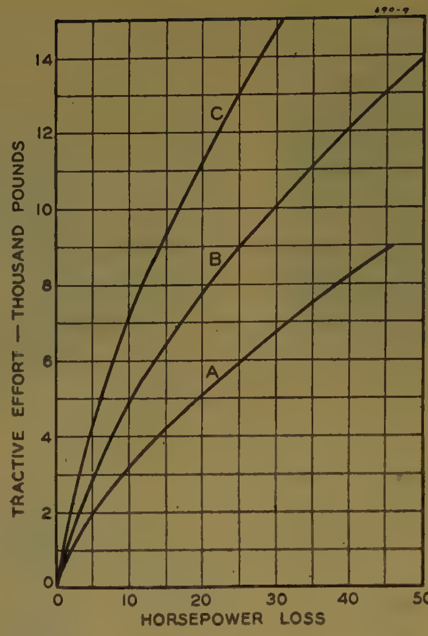


Figure 9. Comparative motor resistance loss curves of traction motors with single-reduction and double-reduction gearing

tractive efforts. Figure 9 shows comparative efficiencies of various motors based on motor resistance losses at 110 degrees centigrade copper temperature, with an assumed two volts for brush drop. Curve *A* is for the older type of single-reduction motors, such as were being built in the period 1925 to 1930. Curve *B* is for single-reduction motors as they are built today. Curve *C* is for the double-reduction motor. All three of these curves are based on the maximum reduction which can be used and still maintain clearance under the gear case as specified by the Interstate Commerce Commission. A study of these curves clearly shows the advantage of the double-reduction motor for locomotives that are required to operate at moderate or low speed and to produce high tractive effort. At the starting point of the locomotive, all of the losses are in resistance. The difference between curve *B* and curve *C* in Figure 9 shows the additional tractive effort that can be produced with the modern double-reduction motor over the modern single-reduction motor for a given horsepower loss. For example, this amounts to 47 per cent increase at 30 horsepower loss.

Universal Use

The modern double-reduction motor is used on a wide range of locomotive sizes. The motor has been applied on locomotives for all classes and types of industrial switching, and also for lightweight railway switching and road service.

A rule-of-thumb measure of the capabilities of the motor is given by these facts: with maximum reduction gearing, it can be applied on 50,000 pounds of locomotive weight per motor and will have a continuous tractive effort rating of 11.7 per cent adhesion, with a maximum permissible speed of 20 miles per hour.

Table I lists the locomotives on which the GE-733 motor is used. This tabulation also shows the number of motors

Table I. Tabulation of Locomotives Equipped With Modern Double-Reduction Traction Motors

Weight, Tons	No. of Motors	Max. Speed	Adhesion Per Cent
Industrial Diesel-electric locomotives			
25	1	20	11.7
45	2	20	11.7
65	4	30	11.5
Railway Diesel-electric locomotives			
44	4	35	14.6
70	4	25	12.5

used per locomotive, the maximum permissible speed, and the continuous adhesion rating of the locomotive. The largest number of these motors have been used on the 25-ton and 45-ton locomotives for industrial service and the 44-ton locomotive for railway service. A limited number of motors have been used on the 65-ton industrial and the 70-ton railway locomotives. In all of these applications, the double-reduction motors have given excellent service.

Determination of Cable Temperatures by Means of Reduced-Scale Models

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Synopsis: A method, using reduced-scale model installations, is developed for the laboratory determination of cable temperatures obtaining in the field. The theory of the method is given, followed by the description of three sets of experiments conducted to show the feasibility of the method. One set of experiments is concerned with a single-conductor cable buried directly in the soil of the model setup, and temperatures obtained are compared with computed values. Another set is concerned with a single-conductor cable in a model duct bank, and sheath and duct temperatures obtained are compared with those of field determinations in an actual field installation. The third set is concerned with a model three-conductor cable in a duct bank, and conductor-sheath temperature differentials are compared with field data. All tests show that the accuracy of the method is sufficient for practical purposes.

THE determination of cable temperatures, under both stationary and transient conditions, is essential for the evaluation of maximum permissible loads. This statement follows from a consideration of the factors that limit the loads, that is, deterioration of the paper-oil insulation and undue stretching and longitudinal movement of the sheath. All these factors are considerably dependent on temperature, so that, indeed, temperature limits the loads.

Determinations of cable (conductor and sheath) temperatures can be and sometimes are carried out in the field. If, however, a wide variety of cable combinations and loading schedules is to be investigated, field tests are time consuming and costly.

A second possibility of obtaining cable temperatures is by calculation. This method, however, is suitable only for very simple arrangements, such as a cable buried directly in the ground. For duct systems of cables, a computation on a rigorous basis cannot be carried out.

A third method of attacking the problem is the use of electrical analogies. This method has been developed by Victor Paschkis[1] and applied successfully to a number of thermal problems. For several cables in duct banks, the necessary electrical equipment would be too involved, so that at present this method cannot be used for this purpose.

The method to be described in this paper was developed for the purpose of providing a more satisfactory means than those just described. As will be seen from the following, the equipment even for involved cable problems is relatively simple and, once the installation is assembled, the time required to solve any specific cable temperature problem is relatively short.

Theory

Thermal problems in a system of cables belong, strictly speaking, to the group of 3-dimensional problems, but, if the length of the cable is at least 10 to 20 times larger than the depth of the cable below ground level (a condition that is nearly always fulfilled), the heat flow along the length can be neglected as compared with the heat flow in radial directions. Thus, cable heat-flow problems can be considered as 2-dimensional.

By examining the equations of heat flow in 2-dimensional cases,[2,3] a dimensional analysis shows that there is a possibility of a scale reduction in such problems, and that the rules according to which this reduction is to be carried out are rather simple. Thus, installations can be reduced in size and yet give the same information concerning temperature distribution as the original large installations. This, briefly, is the principle of the method developed and described in this paper.

The following is a brief presentation of the dimensional analysis mentioned. One of the basic equations[2] in 2-dimensional heat problems is the temperature distribution around an instantaneous cylindrical heat source in an infinite medium, the temperature of which becomes zero at infinity. Let H_0 denote the instantaneously generated heat per unit length of the cylinder with radius a, g the thermal resistivity, D the thermal diffusivity of the medium surrounding the source, T the temperature, t the time, and r the radial distance from the axis of the cylinder. Then

$$T=\frac{H_0 g}{4\pi t}\exp\left[\frac{-(r^2+a^2)}{4Dt}\right]J_0\left(\frac{ira}{2Dt}\right) \qquad (1)$$

where J_0 = Bessel function of the first kind and zero order. In this equation, the several variables occur in two combinations. One is (with l = any characteristic length of the installation)

$$x=\frac{l^2}{Dt} \qquad (2)$$

the other

$$y=\frac{T}{Hg} \qquad (3)$$

where H stands for the heat rate H_0/t and

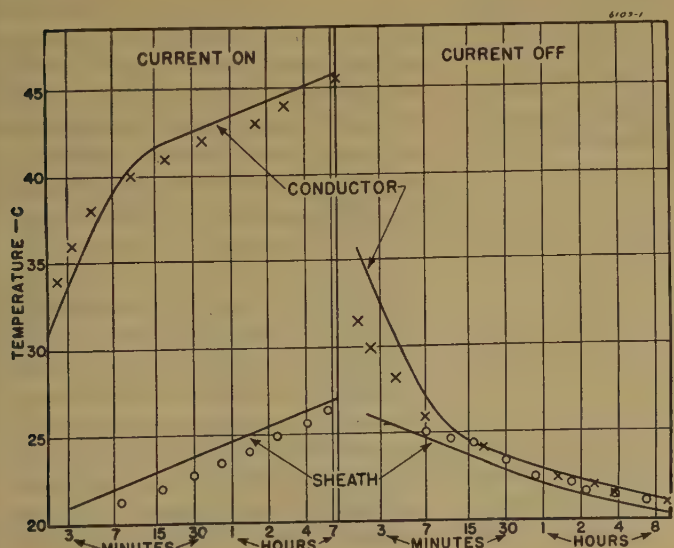

Figure 1. Temperatures of conductor and sheath of one single-conductor cable buried directly in ground of model installations

Eight-hour loading at 120 amperes and subsequent cooling (x and o are measured data; solid curves are calculated)

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has the meaning of heat generated per unit length per unit time. Both x and y are dimensionless, as is shown by the following.

D stands for $l/g\rho c$, with ρ=density and c=specific heat, and has the dimension l^2t^{-1}. Hence, x is dimensionless. Concerning equation 3, T as temperature is dimensionless, the dimension of H (heat per unit length per unit time) is mlt^{-3} (m=dimension of mass), and that of g is $m^{-1}l^{-1}t^3$; consequently, y is dimensionless. As can be seen from equations 2 and 3, equation 1 has the form

$$y=f(x) \tag{4}$$

where f stands for a function of the single argument x, containing no other parameter.

By further examining equations of 2-dimensional heat flow, it will be found that all of them can be reduced to a dimensionless form similar to equation 4. As a further example, a cylindrical space bounded by an inner cylinder of radius a, and an outer cylinder of a radius a' is considered. The inner cylinder generates heat in the amount of H per unit time per unit length; the outer cylinder is maintained at zero temperature. The thermal constants of the medium between the two cylinders are g and D, as before. The temperature of the inner cylinder T at any time t is then given[3] by

$$T=\frac{Hg}{4\pi}\sum_{n=1}^{\infty} k\left[J_0\left(\frac{\rho a}{a'}\right)\right]^2\left[1-\exp\left(-\frac{\rho^2Dt}{(a')^2}\right)\right] \tag{5}$$

where

$$k=\frac{2}{\frac{\pi}{4+n-1}} \tag{6}$$

and, approximately,

$$\rho=2.40+(n-1)\pi \tag{7}$$

the infinite series to be extended from n=1 to infinity.

It can be seen that equation 5 contains all variables in the form of x and y, as before, and, in addition, z; where

$$z=\frac{l}{l'} \tag{8}$$

l and l' denoting two characteristic lengths of the installation. Hence, equation 5 can be written in the general form

$$y=f(x, z) \tag{9}$$

f standing for a function of the arguments x and z only.

From the existence of functional equations such as 4 and 9, the conclusion is drawn that any two installations for which x, y, and z are respectively the same are physically similar. A closer investigation of equations 2, 3, and 8 reveals the conditions of similarity, as follows:

1. Equation 8 shows that a scale reduction of an installation is feasible if all linear dimensions normal to the running direction of a cable or a duct system of cables are reduced in the same ratio. For instance, the ratio a/a' in a concentrical cylindrical case should be the same both in a scale model and an actual cable.

2. If the heat-conducting substances are the same in the two corresponding installations, that is, if D is the same for both, then equation 2 shows that the time t required to attain the same temperature in both cases is reduced by a factor that is equal to the square of the reduction factor of linear dimensions. This can be seen by writing $t=l^2/Dx$ and recalling that x is the same in the original and the model. This indicates the considerable gain in time that may be obtained by carrying out temperature measurements on a reduced-scale installation. It should be added that, while the method covers transient conditions, as shown, it is obviously applicable for the limiting case of $t=\infty$, the steady-state condition. Equation 5, for instance, is valid for the steady-state condition if the exponential is replaced by zero, thus making the second bracket unity. Steady-state conditions are approached, according to equation 2, faster in the reduced-scale installation than in the original.

3. Equation 3 shows the following condition. Since the same materials are used for both the actual installation and the model, g is the same. If, now, the reduced-scale system should yield the same temperatures T as does the original system, the condition that follows from equation 3 is that H, the heat dissipated per unit time and unit length, should be the same in both cases. Disregarding dielectric and sheath losses in obtaining a first approximation, H is given by

$$H=I^2R \tag{10}$$

where I=current in a conductor and R= resistance of the conductor per unit length. Because of condition 1, the cross-sectional area is reduced by the square of the linear reduction factor, and R increases in the same ratio. Thus, in order that H be the same, the current has to be reduced by the same factor as the linear dimensions.

If the reduction factor q be introduced and if quantities referring to the reduced-scale installation be designated by the index r, then condition 1 requires

$$l_r=\frac{l}{q} \tag{11}$$

condition 2 requires

$$t_r=\frac{t}{q^2} \tag{12}$$

and condition 3 requires

$$I_r=\frac{I}{q} \tag{13}$$

The last equation is approximate; since the influence of skin effect varies with the size of conductors, a small correction factor that is available from tables has to be considered. If these three conditions are fulfilled, corresponding temperatures in the model and original installations are the same.

The larger q, the smaller is the model installation. Practical considerations, however, limit its value, and for 3-conductor 24-kv cables, a value in the approximate range of from four to five has been found to be suitable.

Test Results

The theory of scale reduction, as outlined above, does not need experimental proof, as it is based on straightforward mathematics. To what extent it could be put to use could, however, be determined only by experiment, because in building the model installation, some approximation of the theoretical requirements could not be avoided. To mention a few examples: the insulation of the model cable could not be easily built of paper five times thinner than real cable paper, that is, 1 to 0.6 mil thick; the concrete in the duct bank could not be easily built of pebbles reduced five times the original size; and the soil surrounding the duct bank could not easily be maintained at exactly the same moisture content, and hence at the same insulation level, as the soil in the field.

Apart from these approximations, resulting from a compromise because of details of construction, there is another divergence between actual installations and the conditions required by the theory as developed. The latter refers to heat transfer by conduction. This condition generally is fulfilled in cable systems with the exception of the space between the cables and the ducts, in which part of the heat transfer takes place by radiation and convection and for which the scale reduction as developed is not valid. These local processes within the duct play a definite and by no means negligible part in determining the temperature distribution. By means of approximate calculations, it is possible to assess their role in a semi-quantitative manner.

In the following sections, three particular tests, carried out for the purpose of checking the degree of accuracy obtainable in practical use of the theory, are described and the results discussed. One of these tests concerns one single conductor cable buried directly into the ground of the model installation. In this case a comparison of test results is made with calculated results. Such calculations are

feasible in this simple case. The second test refers to a single-conductor cable in a model duct system, and comparison is made with the results of an actual field test in a corresponding installation. The third test refers to a model 3-conductor cable in a model duct system and comparison is made with results of the same full scale field test as in the second test.

All these tests were carried out in a wooden box 15 feet long, 5 feet wide, and 4 feet high filled with soil (a yellow clay obtained from the location of the field installation just referred to). A volume of soil equal to that of the space in the box to be filled was dug in the field and placed in the box by tamping. Only about five cubic feet were left over, so the density of the soil in the box had nearly the original value obtaining in the ground. Moisture determinations revealed that the soil contained originally 17.5 to 20 per cent water. Although a certain decrease of the moisture content with time could not be prevented, lining the box with tar paper, covering it with tarpaulin, and occasional watering of the top kept the drying out of the soil within reasonable limits. For future tests, it is considered advisable to install the model in a room in which the moisture content is artificially maintained at a relatively high level, say 80 per cent.

The thermal resistivity of the soil was measured by means of a cylindrical soil-resistivity cell[4] in the field, and shortly before removal of the sample a value of 72 thermal ohms was obtained. The diffusivity D was obtained as follows. The density ρ was 2.1 grams per cubic centimeter. The clay-to-water ratio was four to one and, since the thermal capacities of clay and water are 0.22 and 1.0 respectively, the thermal capacity of the soil is 0.37. From $D=1/g\rho c$, a value of 4.2×10^{-3} square centimeters per second was obtained for the diffusivity. These values for g and D were used in the calculations.

Single Cable Buried in Ground

This arrangement was selected because the temperature distribution in and around one single conductor cable buried in the ground could be computed theoretically. A comparison between experimental and computed data would indicate the accuracy and reliability of measurements obtained from a small size model installation.

The test cable used was a single-conductor, number 6 American Wire Gauge, paper-insulated cable having an insulation thickness of 9/64 inch, lead sheath thickness of 6/64 inch, and over-all diameter of 0.65 inch. The cable was installed ten inches below the ground surface in the box. The ratio of depth of burial to length of cable being 1 to 18, the error introduced by the end-effect, according to data of S. Whitehead,[4] was between one and two per cent.

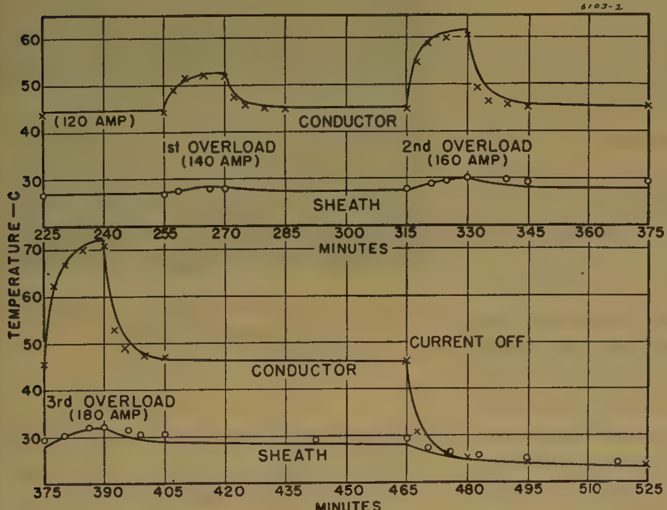

Figure 2. Temperatures of conductor and sheath of one single-conductor cable buried directly in ground of model installation

Overloads of 20, 40, and 60 amperes superimposed upon steady current of 120 amperes (x and o are measured data; solid curves are calculated)

The temperatures of both the conductor and the sheath were measured. The model method permits the copper temperature to be obtained by means of a thermocouple inside the conductor, since it is possible to pull a thermocouple into a narrow channel within the conductor. In this particular case, the central one of the seven conductor strands was pulled out before the test in order to accommodate a thermocouple; another thermocouple was soldered on the lead sheath to obtain sheath temperatures. The temperatures were continuously registered by means of a recording instrument. Thus, it is possible to maintain uninterrupted registration of copper temperatures during the whole load cycle. In contrast to this procedure, copper temperatures in field tests are obtainable only by resistance measurements, necessitating a temporary interruption of the loading.

A current transformer with the necessary control and measuring equipment was provided for the loading of the cable.

With regard to calculation of temperature in this case, equations recommended by S. Whitehead[4] were used. By considering the sheath as the source of heat flow, the basic equation 1 immediately gives the solution in the case of the buried cable. Since the ground surface is an isothermal plane, the field of an image sink above ground, added to the field of the real source, constitutes the solution of the problem. For values of $a^2/4Dt$ that are small as compared to unity, the solution for the sheath temperature T_s becomes

$$T_s=\frac{Hg_s}{4\pi}\left[-Ei\left(-\frac{a^2}{2Dt}\right)+Ei\left(-\frac{b^2}{Dt}\right)+\exp\left(-\frac{a^2}{2Dt}\right)\log 2\right] \quad (14)$$

where g_s is the thermal resistivity of the soil, Ei stands for the exponential integral,[5] and b denotes the depth of the cable below ground, the other symbols being the same as in equation 1.

The rise of the copper temperature above sheath temperature can be computed by considering both the thermal capacity C_c and the thermal resistance G_c of the cable per unit length as lumped elements, in which case the solution is an exponential. This solution, although approximate, is sufficiently accurate for the purpose at hand. The copper temperature is then given[4] by

$$T_c=HG_c\left[1-\exp\left(-\frac{2t}{C_cG_c}\right)\right]+T_s \quad (15)$$

T_s being known from equation 14. The factor 2 in the bracket is approximate: it accounts for the fact that the center of the cable is heated instantaneously, while the periphery is heated across the resistance G_c; in the average, a resistance of $G_c/2$ is operative.

The numerical values used for the two constants G_c and C_c were evaluated from the geometry of the cable and the known characteristic data of each cable constituent. G_c is given by

$$G_c=\frac{g_c}{2\pi}\log\frac{r_s}{r_c} \quad (16)$$

with g_c = thermal resistivity of insulation, r_s = inside radius of sheath, and r_c = radius of conductor. For g_c a value of 550 thermal ohms was taken. With $r_s=0.232$ inch and $r_c=0.092$ inch, the value

obtained for G_c is 81 ohms or 81×10^{-7} centimeter-gram-second units.

The thermal capacity was computed by adding the values for copper, insulation, and lead. The specific heats are respectively: 0.42×10^7, 1.8×10^7, and 0.147×10^7 centimeter-gram-second units, resulting in thermal capacities per unit length of 0.42×10^7, 1.98×10^7, and 1.78×10^7 centimeter-gram-second units, giving a total for $C_c=4.18\times10^7$ centimeter-gram-second units.

The steady-state temperatures are given by the well-known formulas:

$$T_s=\frac{Hg_s}{2\pi}\log\frac{2b}{a} \tag{17}$$

and

$$T_c=HG_c+T_s \tag{18}$$

Since the temperature of the conductor and, consequently, its resistance varies during a load cycle, in order to maintain a constant heat loss H, upon which the theoretical equations used for calculation are based, the current must be adjusted occasionally. This adjustment is possible when the copper temperature is registered continuously. The currents reported in the subsequent parts of this paper are the values applied at 20 degrees centigrade copper temperature and at which the conductor resistance was 15.4×10^{-6} ohms per centimeter.

The first test was carried out with a load of 120 amperes corresponding to $H=6.8$ watts per foot. Loading continued for about eight hours, after which the current was switched off. Both the heating and the cooling curves—temperature versus time—are shown on Figure 1 for the conductor as well as for the sheath. The time is plotted on a logarithmic scale. The crosses and circles indicate measured data; the solid lines were computed from equations 14 and 15, using the numerical values given.

The agreement between calculated and measured data is satisfactory, the discrepancies being generally one to two degrees centigrade, except the copper temperature within five minutes after switching off the current, where the divergence is up to four degrees centigrade. The amount of disagreement between computation and test is partly caused by the fact that the equations, although theoretical and without arbitrary constants, contain, as explained above, some unavoidable approximations. Part of the disagreement, however, is certainly caused by deficiencies of the equipment (end-effect, boundary layer between sheath and soil, drying of the soil, time lag in temperature recorder, and so forth). But

Figure 3. Details of assembly of model ducts in model trench

since a disagreement of less than two degrees centigrade is of no great practical significance, and since the cooling period is of relatively little interest to the practical cable engineer, who is not greatly concerned with the precise rate of temperature decrease after the temperature has reached its maximum, the method is considered satisfactory. For higher accuracy some of these deficiencies might be eliminated: for instance, the use, instead of moist soil, of a suitable dry and stable powdered material having the same thermal resistivity is being considered.

A second test (Figure 2) shows the temperature changes during and after short overloads. A current of 120 amperes was first maintained for about four hours (255 minutes), after which three overload cycles of 60 minutes duration, with overloads on for 15 minutes each, were applied. The total overload currents were respectively 140, 160, and 180 amperes (values reduced for 20 degrees centigrade copper temperature). The corresponding heats dissipated are 9.2, 12.0, and 15.2 watts per foot. The degree of agreement between test results and calculation is the same (deviation generally within two degrees centigrade) as in the first test. The agreement during the temperature maxima, which are of particular interest from a practical standpoint, is rather good. These findings are encouraging with regard to the practical application of a model installation (see conclusions at end of paper).

Single Conductor Cable in Duct Bank

This test was carried out in order to show whether the model was satisfactory in a case involving ducts. Since computa-

Figure 4. Completed model duct bank prior to pouring of concrete

tion for such a case is not possible, a set of field tests conducted by The Detroit Edison Company was duplicated in the model setup to determine the degree of agreement between the two.

It is planned for all tests of this type in the future to use reduced scale replicas of those cables that are involved in the actual field installations in question. At the time of carrying out the test to be described in this section, such model cable was not available, and the same single conductor cable that was used in the preceding test was installed in the model duct bank. In spite of this difference, sheath and duct temperatures, apart from the initial transients, in the model and the field should be the same. Because of the difference in the cable structures, however, copper temperatures in the field and in the model were not comparable in this particular test, and, as they are not essential in answering the specific question concerning the role of ducts in scale model tests, they are not given here.

A fiber-conduit–concrete duct bank for 12 4-inch ducts (3 wide by 4 high; $22^1/_2$ by $28^1/_2$ inches) was reproduced in a reduced size model. The linear reduction

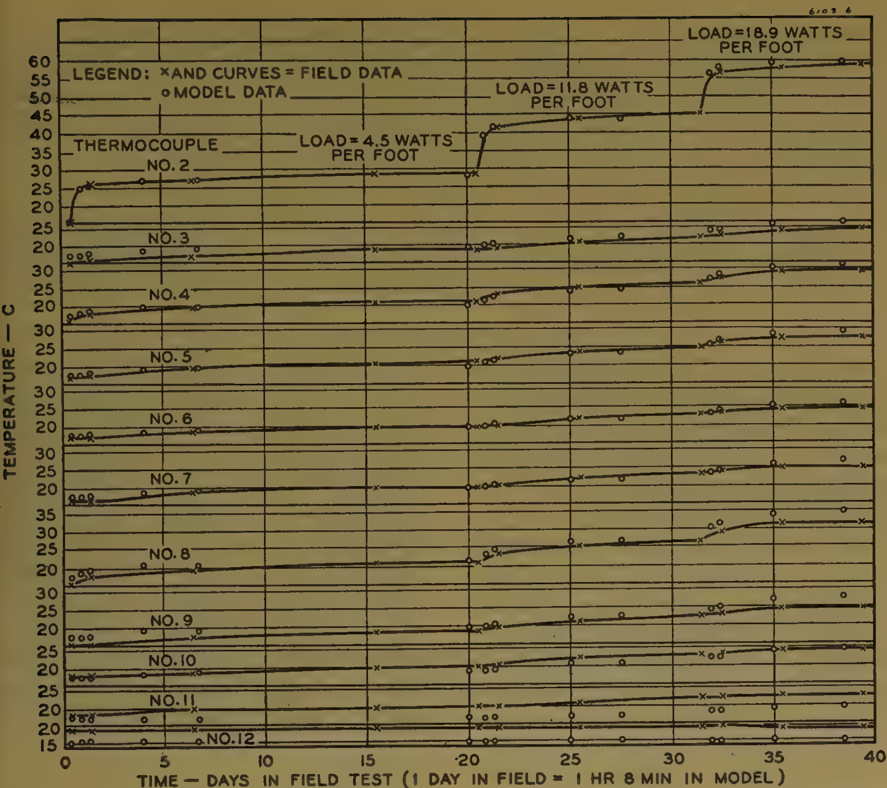

Figure 6. Temperatures of a test cable, ducts, and ground at three different indicated loads for both field and model installations

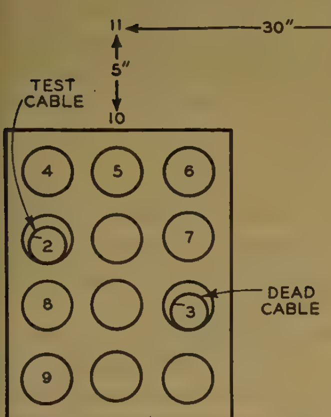

Figure 5. Diagram (not to scale) of model duct with number and location of thermocouples

factor q (see equation 11) was chosen as 4.6. This figure was selected in order to enable number 8 American Wire Gauge stranded wire to be incorporated in the model 24-kv 350-Mcm round conductor cable in the subsequent tests. The conductor diameter of the 24-kv cable is 0.68 inch, that of the number 8 wire 0.147 inch, the ratio of the two being 4.6. With $q=$ 4.6, the inside diameter of the model ducts was 7/8 inch, and the wall thickness was 1/16 inch. Three-foot lengths of bakelite-impregnated linen tubing, joined by short pieces of 1-inch inside diameter tubing as sleeves, were used as the conduit. The thermal resistivity of this material is comparable with that of fiber conduit as used in the field. Scale model concrete spacers between the ducts were used. A scale-model trench with a cross-sectional area of 4.9×6.2 inches was dug in the soil contained in the wooden box mentioned previously, and the duct bank was assembled in the trench. The manner in which the ducts were assembled is shown in Figure 3.

Two bakelite plates attached to the end walls of the box fixed the position of the ends of the duct bank. Pebble-free concrete containing sieved sand mixed with fly ash and having a two to one sand-to-cement ratio, was poured into the trench. The soil was tamped down on top of the duct bank, the top of the concrete being $11^1/_2$ inches below the top of the soil (corresponding to four feet four inches below ground level in the field). The completed assembly, prior to pouring the concrete and refilling the soil, is shown in Figure 4.

The ends of the bank were protected from air currents by close-fitting wooden boxes attached to the outside of the main box.

The arrangement of the cables and the location and number of thermocouples, as used in the field test and in the model, are shown in Figure 5. The length of the test cables was 17 feet. It should be noted that curve 2 (thermocouple 2) refers to the sheath of the test cable (thermocouple 1, as used in the field test was not reproduced in the model); curves 3 to 9 inclusive refer to the air temperature in the center of the various ducts; and curves 10 to 12 inclusive refer to the soil.

The results of the test are shown in Figure 6. The crosses and the curves drawn through them show the temperatures, in degrees centigrade, indicated by the 11 thermocouples used in the field tests; the circles show the data obtained with the model. The 40-day field test took 45 hours in the model setup (the time reduction factor from equation 12 is $4.6^2=21.2$).

Since it was realized that the ducts in this particular test were wet, some water was introduced in the duct containing the test cable in the model. A total of 200 cubic centimeters was introduced in order to insure a conducting path between cable and duct wall. This point is discussed in detail later.

Three different consecutive loads were applied. According to the requirements of the model theory, the heat input per unit length of cable has to be the same in the model as in the field. The three loads applied were 4.5, 11.8, and 18.9 watts per foot, respectively.

In studying Figure 6, it can be seen from curve 12 that no attempt was made to duplicate in the model setup the ambient temperatures recorded in the field; hence, the same disagreement prevails also for curve 11. The agreement for the sheath, the various ducts, and the soil just outside of the bank (shown by curve 10) is reasonably good, with a deviation of no more than about three degrees centigrade.

Additional Tests on Single-Conductor Cable in Duct Bank

While the results shown in Figure 6 answer the chief question, namely, whether the model method works in the case of ducts, it was deemed advisable to supplement the results by a few additional tests. As mentioned previously, one of the main causes of doubt with regard to the feasibility of the model test in the case of ducts is that heat is transferred not only by conduction, but also partly by radiation and convection. The scale reduction theory, however, took into consideration conduction only. If, now, the two other mechanisms play only a minor role in ducts, the error in the model test is relatively small. While the satisfactory answer of the main test (Figure 6) answers this question in a favorable sense, it was thought advisable to strengthen the confidence in the model method by additional tests directed toward showing spe-

cifically whether the errors introduced by radiation and convection in a duct are appreciable or not.

Radiation

It is useful in this connection to consider the Christiansen equation, which controls radiation from a convex body in an enclosure[6] and which, applied to the present problem, reads approximately:

$$H_r = 2\pi\sigma a \frac{(T_s^4 - T_d^4)}{\left(\dfrac{1}{\epsilon_s} + \dfrac{1}{\epsilon_d} - 1\right)} \tag{19}$$

where H_r = heat radiated per second, σ = Stefan-Boltzmann constant = 5.7×10^{-5} absolute units, a = radius of sheath, T_s and T_d = sheath and duct temperatures, in absolute units, and ϵ_s and ϵ_d = emissivities of sheath and duct. The emissivities of a tarnished lead surface and a fiber surface are 0.56 and 0.93, respectively.

In order to assess experimentally the role of radiation in the model setup, the following procedure was adopted. From equation 19 it can be seen that ϵ_s has a marked influence on the radiation. Thus by varying the emissivity of the cable surface, all other conditions being left unchanged, it is possible to determine the magnitude of the role of the radiation in a duct. The variation of ϵ_s was carried out by coating the cable with flat black, thus producing a surface with an emissivity of 0.97. The factor

$$\frac{1}{\dfrac{1}{\epsilon_s} + \dfrac{1}{\epsilon_d} - 1}$$

is then changed from 0.54 to 0.90, that is, increased by 66 per cent.

The test described in the preceding section, "Single Conductor Cable in Duct Bank," was repeated with the black sheath, and in this manner sheath temperatures, for the same watts input, were obtained with the cable coated black as were obtained with the cable in its original condition. Results, obtained at identical loads and times for the two surface conditions, are shown in Table I. The duct was dry in these tests; hence the temperatures are higher than shown in Figure 6. It can be seen that, considering the differences in the respective starting temperatures, the increases in temperature are practically identical for the two different surfaces.

The results show that a substantial change in the radiation does not cause a measurable change of the sheath temperatures in the model setup. This conclusion, while reassuring from our particular standpoint, does not, however, prove that there is equally little change between sheath temperatures in the field and the model, in case the radiation resistance in the two is not the same. An estimate, based on equation 19, will provide this missing information for the steady-state condition.

Table I. Sheath Temperatures, Degrees Centigrade, for Different Sheath Surface Conditions

	Load = 4.5 Watts Per Foot		Load = 11.8 Watts Per Foot		Load = 18.9 Watts Per Foot	
Time, Hours	Plain Lead	Flat Black	Plain Lead	Flat Black	Plain Lead	Flat Black
0	17	20	35	37	63	62
1	33	35	59	60	84	82
2	33.5	36	60.5	61	86	85
3	34	37	60.5	62	87	85.5

Since equation 19 contains the radius a, the radiation resistance of the duct space in the model is 4.6 times that in the field, while the thermal resistance remains unchanged in a scale-reduced model. If the total heat flow is the same in both cases, the sheath temperature in the model, because of the higher radiation resistance, will be somewhat higher than in the field. It can be shown, however, that a very little increase, of the order of one to two degrees centigrade, is sufficient to satisfy the equations of heat flow in the two cases.

All data in equation 19, with the exception of T_d, are known. The latter was determined, at least approximately, by introducing a thermocouple into the duct containing the test cable and taking duct air temperatures. For a load of 11.8 watts per foot, $T_d = 35$ degrees centigrade (or 308 degrees Kelvin) was obtained, and it is probably correct to say that the same (or nearly the same) duct temperature exists in the field, too. With a cable diameter of three inches and $T_s = 43$ degrees centigrade = 316 degrees Kelvin, equation 19 yields for H_r in the field 2.2 watts per foot, which is about 20 per cent of the total heat. Neglecting convection (see the following), the balance of heat, namely 9.6 watts per foot is transferred by conduction.

The general equation for equilibrium conditions of temperature and heat flow in the duct is

$$H = A(T_s - T_d) + B(T_s^4 - T_d^4) \tag{20}$$

where A and B are the conductances caused by heat conduction and radiation (the latter given by equation 19), and in the present case (for the field):

$$11.8 = 1.2\,(T_s - 308) + 2.2 \times 10^{-9} \times (T_s^4 - 9 \times 10^9)$$

and this equation is satisfied with T_s = 316 degrees Kelvin = 43 degrees centigrade. In the model, A remains the same, but B is reduced by a factor of 1/4.6. Hence, for the model:

$$11.8 = 1.2(T_s - 308) + 0.48 \times 10^{-9} \times (T_s^4 - 9 \times 10^9)$$

and it can be shown that a value T_s = 317.3 degrees Kelvin = 44.3 degrees centigrade satisfies this equation. This value is only 1.3 degrees centigrade higher than that in the field, and this difference is well within the reported deviation of 3 degrees centigrade. This conclusion is valid for wet ducts, which constitute the majority of existing ducts. For dry ducts the error caused by radiation is certainly larger.

Convection

A certain amount of heat is transferred across the duct space by means of free convection. There are empirical equations,[1,6] that permit approximate computations. H_c, the heat transferred per unit length by convection from a pipe to air, is given by

$$H_c = ka^{3/4}(T_s - T_d)^{5/4} \tag{21}$$

where k = a constant depending on the ratio of cable radius to duct radius; for the present case, $k = 2.7 \times 10^3$ absolute units.

In order to check the influence of convection on the sheath temperature, the free convection normally active around the sheath was increased by forced convection. This was obtained by forcing nitrogen longitudinally through the duct. If V denotes the rate of flow of the nitrogen, ρ_N its density, c_N its specific heat, and ΔT the temperature differential along the duct, then the heat removed per unit time from the sheath by way of forced convection, H_{fc}, is given by

$$H_{fc} = V\rho_N c_N \Delta T \tag{22}$$

V and ΔT were measured, and H_{fc} calculated from equation 22. The rate of flow, V, was adjusted so that H_{fc} equalled approximately twice the value of H_c (known from equation 21) times the length of the duct. The sheath temperatures measured before, during, and after applying the forced flow gave an indication of the influence of convection on the temperature distribution.

Table II shows the data obtained on two loads: 11.8 and 18.9 watts per foot,

Table II. Sheath Temperatures, T_s, for Different Conditions of Convection

	Load = 11.8 Watts Per Foot		Load = 18.9 Watts Per Foot	
	Free Convection Alone, 5 Watts	Forced Convection Superimposed, 8.5 Watts	Free Convection Alone, 8.5 Watts	Forced Convection Superimposed, 19 Watts
Before forced convection	59	—	84	—
During forced convection	—	59.5	—	83
After forced convection	60	—	83	—

with the duct dry. T_s for the two loads is 59 and 84 degrees centigrade; T_d is 37 and 51 degrees centigrade respectively.

It can be seen that the influence of a forced convection, amounting in terms of heat loss to about double that for free convection, is practically negligible. As before, in the case of radiation, this result refers to the model only and does not give specific information as to a divergence between field and model data caused by a possible difference in convection. This difference, however, can be estimated from equation 21.

Considering the medium load, 11.8 watts per foot, the value of H_c, the heat dissipated by way of convection, for the field and the model respectively, is 0.3 and 0.1 watt per foot, that is three and one per cent of the total. In contradistinction to radiation, which amounts to appreciable percentages of the total heat dissipated—without, however, affecting correct duplication of the field test by the model—convection amounts to only a small percentage of the total heat dissipated. The effect of this small amount, being reduced in a ratio of one to three in the model as compared with the field, has certainly no influence upon the duplication of the sheath temperatures.

Water in Ducts

As a last additional test, the question of the amount of water that is required to make a wet duct out of a dry duct was investigated. In comparing sheath temperatures shown in Table I (dry duct) with those shown for the model test in curve 2 of Figure 6 (wet duct), it can be seen that a dry duct results in sheath temperatures higher than those in a wet duct. It was important to find out how much water should be introduced in the model duct in order to approximate field conditions. From physical considerations it appeared that a certain minimum amount is required in order to establish a conducting path of water between the cable and the bottom of the duct wall and that further additions of water should have little effect. Table III gives the results of a test, referring to a load of 18.9 watts per foot, in which increasing amounts of water were added to the duct containing the test cable. It can be seen that for the model, small amounts of water (ten cubic centimeters) were sufficient to bring the temperature of the sheath from 83 degrees centigrade down to 65 degrees centigrade. It appears that ten cubic centimeters is sufficient to cause an appreciable drop in temperatures and that 50 cubic centimeters is sufficient to insure the formation of a complete conducting path along the length of the model duct and to bring the temperature level down to 60 degrees centigrade, which was the level as measured in the main test and also in the field. Further additions of water change the temperature only slightly. The equivalent amount of water present in a wet duct in the field is then at least $10\times21.2/15=14$ cubic centimeters per foot.

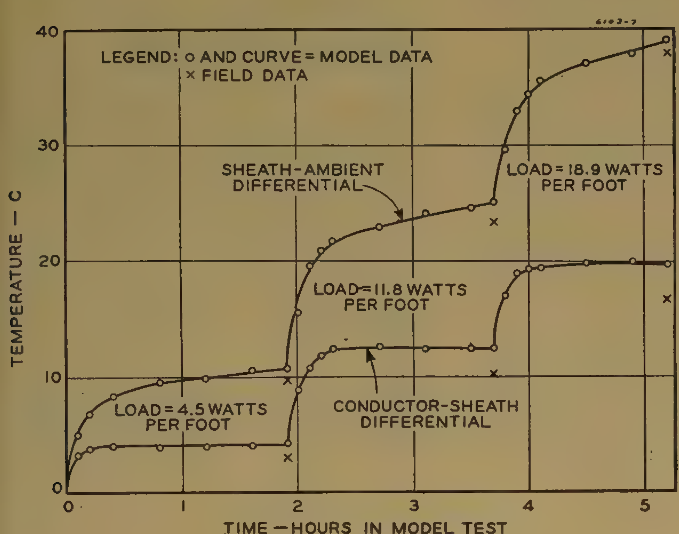

Figure 7. Temperature differentials between sheath and ambient, and conductor and sheath for a 3-conductor model cable

Installation and load schedule is the same as indicated in Figures 5 and 6. Field data for the conductor are equilibrium values

Model 3-Conductor Cable in Duct Bank

The tests described in the two preceding sections were carried out with a single-conductor cable, and accordingly conductor temperatures were not comparable with those in the field because the field tests involved 3-conductor cables. In order to fill this last gap in the research, it was necessary to have a model of the 3-conductor 350,000-circular-mil cable used in the field. Through the courtesy of Doctor R. J. Wiseman, the Okonite Callender Cable Company manufactured such a cable according to developed specifications.

The cable was constructed according to the following design. The conductor was number 8 American Wire Gauge wire having an over-all diameter of 0.147 inch, which is, as pointed out previously, 4.6 times less than the conductor of the high-voltage cable. One of the three conductors was a copper tube of an outside diameter of 0.156 inch and a wall thickness of 0.032 inch; the cross-sectional area of the tube was nearly identical with that of the stranded wire. The purpose of the tube was to accommodate the thermocouple for conductor temperature measurements. The insulation

Table III. Sheath Temperatures, Degrees Centigrade, for Increasing Amounts of Water in Duct

Load = 18.9 Watts Per Foot

Amount of Water, Cubic Centimeters, in 15-Foot Duct Containing Test Cable	Sheath Temperature, Degrees Centigrade, of Test Cable
0	83
10	65.5
20	57 (?)
30	67
40	65.5
50	62
60	60
70	61
80	62
100	60.5
120	59
140	58
200	59.5

thickness was $0.31/4.6 = 0.068$ in. The outermost layer was metallized paper. In order to compensate for the larger outside diameter of the tube, one layer of paper was omitted from its insulation. Fillers of suitable size and a binder tape made of 4-mil copper were used. (A copper tape of reduced thickness, say, 2-mil, is more advisable.) The lead sheath was 2/64 inch thick, and the over-all diameter obtained was 0.69 inch, which is close to the theoretically correct 0.68 inch, that is, 3.1/4.6.

The single-conductor cable used as test cable in the last test described (Figure 6) was then replaced by this model cable, and the same load cycle schedule applied, except that for the sake of simplification the duration of the cycles was not more than about two hours each. This was sufficient in this case, since the main interest of this test centered around the sheath-conductor temperature differential, and this reached its final value in about that time.

The results of this test are reproduced in Figure 7. Both sheath-to-ambient and conductor-to-sheath differentials are shown. The corresponding field data for the conductor refer to equilibrium values as obtained at the end of each load period.

While the sheath temperatures are in rather good agreement with those in the field, the discrepancy for the conductor-sheath differential is somewhat larger, up to three degrees centigrade. This difference is partly inherent, inasmuch as the insulation structure, and therefore the thermal resistivity, for a real and a model cable cannot be made strictly identical. Part of the difference, however, is caused by uncertainties and difficulties in the measurement of copper temperatures in the field (obtained from the electrical resistance of the conductor). A deviation of up to three degrees centigrade, nevertheless, is not objectionable from a practical standpoint.

In closing, a few remarks concerning end-effect are in order. Since these tests were carried out with the idea of later developing the method into a practical one, it was tried to keep it as simple as feasible; no attempt was made, therefore, to compensate for a possible end-effect. In order to ascertain whether or not an appreciable error is introduced on this account, the last test with the model cable was used also for the measurement of the magnitude of this effect. A thermocouple was inserted into the tube, about six inches inside the end walls of the box. At the end of the first load period, the end was cooler by one degree centigrade than the center; at the end of the second load, the end was warmer by two degrees centigrade than the center; and at the end of the third load, this difference increased to six degrees centigrade.

An approximate calculation shows that these differentials are not likely to lead to an appreciable error. From the dimensions of the cable and the thermal resistivity of copper, the thermal conductance of the three conductors from both ends to the center is found to be 0.0013 watt per degree centigrade. For the worst case (the third load) the heat that flows toward the center is only 0.008 watt. Even with the conservative estimate that this heat is dissipated along a length of, say, only one inch, for which the load is 1.6 watts, the error is only one-half per cent, and thus is negligible.

Conclusions

The present research shows that a reduced scale model installation can be used successfully for determining cable temperatures and hence maximum permissible loads, both normal and overload. Such an installation is meant to be used as a laboratory tool, just as any other measuring device. Its chief use probably will be found in cases of several cables in a duct bank, each cable having a different prescribed load schedule. By reproducing any desired cable and load combination in the model, a reliable answer to any specific problem can be obtained in a relatively short time.

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Parallel Circuits in Servomechanisms

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AN excellent way to describe a servomechanism is to picture it as a device which acts with power on the difference, or error, between a desired quantity and an actual quantity. It acts in such a way as,

1. To make the error approximate zero in the steady state.

2. To limit peak errors and the error-time integral when the output quantity is following a typical random input quantity.

3. To recover from sudden disturbances with adequate speed of response and degree of stability.

A block diagram of a group of components is illustrated in Figure 1 in which the foregoing interpretation is very powerful. Much has been written on the analysis and synthesis of such systems. Several recently published or released works have described powerful analytical tools used in the design of linear servomechanisms of this type.[1–4,7,8]

This paper deals in particular with a class of servomechanisms which has not received adequate treatment thus far. Many servomechanisms, as in Figure 2, can be described accurately only when a component is included in the feed-back path. At times this is desired, and compensating devices are used to aid the performance of the servomechanism. At other times, undesirable characteristics of the measuring instruments introduce delay in the transmission of signals back to a point where they are to be compared with the desired quantity. The signals then used to excite the control amplifier are no longer the error between desired and actual quantities.

In cases where there is a modification of the controlled quantity previous to comparison with the desired quantity, no place exists in the circuit where the actual error can be measured. This is most likely to be true when the servomechanism is not in steady-state operation. The ratio of output to error as a function of frequency when the input is varying sinusoidally is, in these cases, not a simple function of any one of the components or any of the components in cascade. This will be developed mathematically as a part of the linear treatment of servomechanisms of the type represented in Figure 2.

General Analytical Techniques

A mathematical treatment of a dynamic system as complicated as a servomechanism is difficult on the basis of classical mathematics. The differential operator p is used here to represent an operator which will be replaced by $j\omega$ when the steady-state sinusoidal characteristics of the system are to be considered, or by d/dt when the differential equations are to be considered.

The general analytical techniques described here are, basically, simple extensions of previous work.[1,2] In electric network theory, the consideration of all circuits on the basis of impedance characteristics to the exclusion of admittance characteristics would be foolish. Similarly in a study of servomechanisms it is often desirable to work with component or system functions which are ratios of input to output instead of their reciprocals.

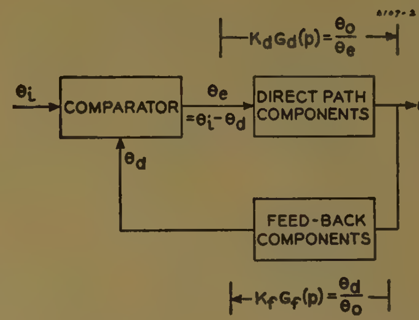

Figure 2. Block diagram of a servomechanism where the components in the direct circuit are paralleled by components in a feed-back circuit

Components of servomechanisms (such as amplifiers, thermocouples, tachometers, and motors) can be considered as analogous to a four terminal network whose input and output are related by a differential equation describing its behavior. In this paper a system function will be defined as a nondimensional ratio of output to input. A component will have a system function, and a combination of components will have a system function. It is, therefore, necessary to describe carefully what a given system function includes. It is also necessary to describe the operator used in expressing the system function. If the operator is a pure imaginary frequency, then a ratio of θ_o to θ_i will be written

$$\frac{\theta_o}{\theta_i}(j\omega)$$

The more general operator p is implied by the simple expression θ_o/θ_i, instead of always writing $\theta_o(p)/\theta_i(p)$. Similarly $\theta_i(p) - \theta_o(p)$ will be written $\theta_i - \theta_o$.

A reciprocal system function will be a

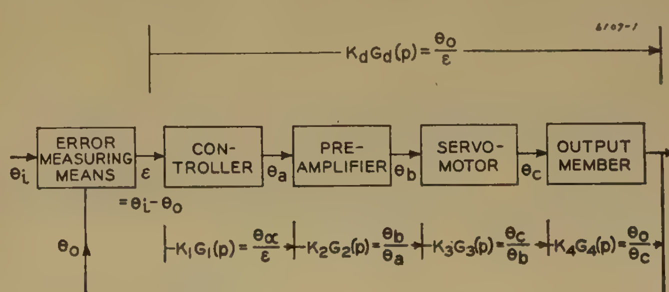

Figure 1. Block diagram of a servomechanism where the components may be considered as elements in a cascade circuit

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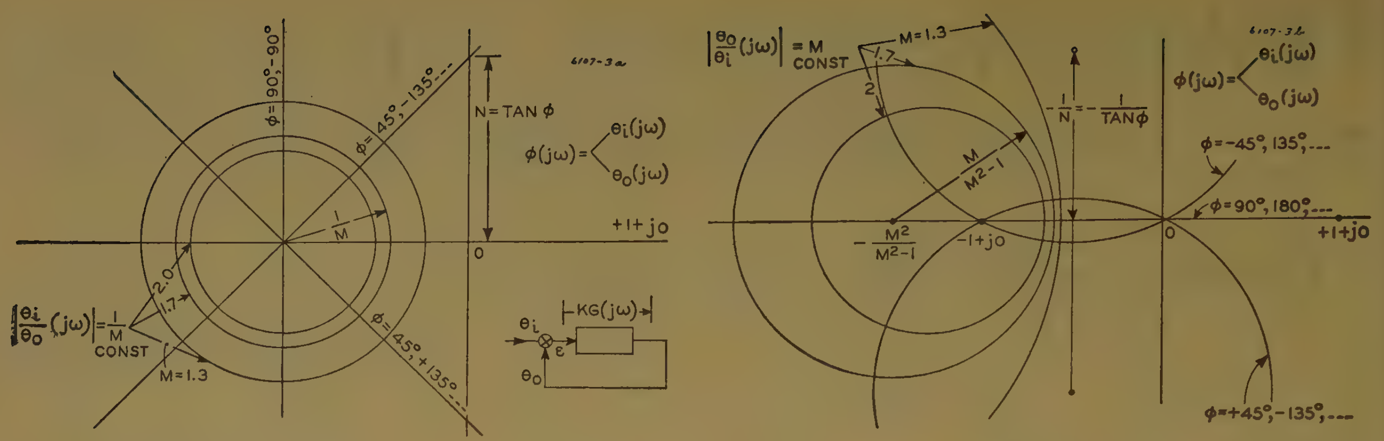

nondimensional ratio of input to output. A component will have a reciprocal system function, and a combination of components will have a reciprocal system function.

The use of reciprocal system functions allows a separation of the component characteristics not otherwise realizable, which leads to greater simplicity of analytical treatment. With reference to Figure 1

$$\frac{\theta_o}{\epsilon}(j\omega) = K_dG_d(j\omega) = K_1G_1(j\omega)K_2G_2(j\omega) \times K_3G_3(j\omega)K_4G_4(j\omega) \quad (1)$$

and

$$\frac{\theta_o}{\theta_i}(j\omega) = \frac{K_dG_d(j\omega)}{1+K_dG_d(j\omega)} \quad (2)$$

Inversion of equation 1 leads to

$$\frac{\epsilon}{\theta_o}(j\omega) = \frac{1}{K_dG_d(j\omega)} \quad (3)$$

and inversion of equation 2 leads to

$$\frac{\theta_i}{\theta_o}(j\omega) = 1 + \frac{1}{K_dG_d(j\omega)} \quad (4)$$

and demonstrates that the system may be studied by a treatment of the function $K_dG_d(j\omega)$, or by a treatment of the reciprocal function

$$\frac{1}{K_dG_d(j\omega)}$$

It does not matter which technique is used, because the effect of any one component, for instance $K_1G_1(j\omega)$, may be studied by a vector product manipulation on a polar plot.

In one important way, however, the use of reciprocal system functions simplifies the problem of synthesis of simple cascaded systems. In a complex plane, in which the vector from the origin at $\omega = \omega_1$ is

$$\frac{1}{K_dG_d(j\omega_1)}$$

loci for

$$\left|\frac{\theta_i}{\theta_o}(j\omega)\right| = \text{constant}$$

are circles, concentric about the $-1 + j0$ point with a radius equal to the magnitude. Loci of constant angle between input and output are straight lines which pass through the $-1 + j0$ point and a point on the imaginary axis equal to the tangent of the angle. These loci are shown in Figure 3, where they may be contrasted with similar loci as they appear on a plane where the vector from the origin at $\omega = \omega_1$ is $K_dG_d(j\omega_1)$.

A simple demonstration of the nature of the loci of constant

$$\left|\frac{\theta_i}{\theta_o}(j\omega)\right|$$

is to set the radius vector

$$\frac{1}{K_dG_d(j\omega)} = R + jX$$

and find the analytical geometric equation for constant

$$\left|\frac{\theta_i}{\theta_o}(j\omega)\right|$$

Substituting in equation 4

$$\frac{\theta_i}{\theta_o}(j\omega) = 1 + R + jX \quad (5)$$

from which

$$\left|\frac{\theta_i}{\theta_o}(j\omega)\right| = \sqrt{(R+1)^2 + X^2} = \frac{1}{M} \quad (6)$$

where the constant M has been defined so as to be consistent with Hall.[1,2] Equation 6 represents a circle whose center is at $-1 + jO$ and whose radius is $1/M$.

The manner in which loci of constant

$$\left|\frac{\theta_i}{\theta_o}(j\omega)\right|$$

and loci of constant angle between θ_i and θ_o appear on the complex plane of

$$\frac{1}{K_dG_d(j\omega)}$$

Figure 3. Loci of constant magnitude of the ratio of the output quantity to the input quantity of a servomechanism and loci of constant phase angle between the output quantity and the input quantity of a servomechanism as they appear on two complex planes

A. Plane of $1/KG(j\omega)$. A complex plane where the radius vector is the complex ratio of the error of the servomechanism to the output quantity

B. Plane of $KG(j\omega)$. A complex plane where the radius vector is the complex ratio of the output quantity to the error of the servomechanism

is important only when these loci are used as design criteria for the servomechanism. In design work the degree of resonance often is limited by specifying a maximum allowable value of M. In terms of the degree of stability of servomechanism transient response this is a rough approximation at best. This is unfortunate because specifications for the speed of response and degree of stability of servomechanisms usually are given in terms of their transient response.

The transient response is related mathematically to the frequency response by the Fourier integral.[5] A perhaps oversimplified interpretation of this relationship is used in specifications of servomechanisms. Such specifications limit the amplification, by resonance effects, of the input and demand a sizable output magnitude at high frequencies. For complicated systems, however, this gives the designer something tangible toward which to work, as well as a good approximate prediction.

The use of the complex plane of

$$\frac{1}{K_dG_d(j\omega)}$$

permits a quick determination of resonance by noting the proximity of the locus to the $-1 + j0$ point. This advantage is, of course, not fundamental. It

is apparent in the solution of numerical problems.

To return to the more complicated system represented in Figure 2: the ratio of output to input is

$$\frac{\theta_o}{\theta_i}(j\omega)=\frac{K_dG_d(j\omega)}{1+K_dG_d(j\omega)K_fG_f(j\omega)} \tag{7}$$

and the ratio of output to error (defined as $\theta_i - \theta_o$) is

$$\frac{\theta_o}{\epsilon}(j\omega)=\frac{K_dG_d(j\omega)}{1+K_dG_d(j\omega)K_fG_f(j\omega)-K_dG_d(j\omega)} \tag{8}$$

This is a hopeless tangle to unscramble when an alteration in either $K_dG_d(j\omega)$ or $K_fG_f(j\omega)$ is needed. However the inversion of equation 7 leads to

$$\frac{\theta_i}{\theta_o}(j\omega)=\frac{1}{K_dG_d(j\omega)}+K_fG_f(j\omega) \tag{9}$$

and the inversion of equation 8 leads to

$$\frac{\epsilon}{\theta_o}(j\omega)=\frac{1}{K_dG_d(j\omega)}+K_fG_f(j\omega)-1 \tag{10}$$

Both equation 9 and equation 10 make possible a separation of the frequency characteristics of the two essential components in the system. Equation 9 and equation 10 indicate a vector addition of the characteristics of the components. The result is obtained with comparative ease and leads quickly to a qualitative idea of the effect of a change in one or the other of the two parallel components.

It would seem from examination of equation 9 and equation 10 that the ratio

$$\frac{\theta_i}{\theta_o}(j\omega)$$

might be plotted instead of

$$\frac{\epsilon}{\theta_o}(j\omega)$$

since the criteria of resonance concerns

$$\frac{\theta_i}{\theta_o}(j\omega)$$

directly. If this is done then loci of constant

$$\left|\frac{\theta_i}{\theta_o}(j\omega)\right|$$

are obviously circles about the origin.

When problems are actually being solved, it often will be convenient to plot on the same sheet of paper a locus of

$$\frac{\epsilon}{\theta_o}(j\omega)$$

and a locus of

$$\frac{\theta_i}{\theta_o}(j\omega)$$

With reference to equation 9,

$$\frac{1}{K_dG_d(j\omega)}$$

may be plotted as the radius vector, and the resonance of the system may be studied by considering the locus to be

$$\frac{\epsilon}{\theta_o}(j\omega)$$

where $K_fG_f(j\omega)$ is always unity. For this purpose circles may be constructed about the $-1+j0$ point. Then, on the same sheet of paper a locus of $K_fG_f(j\omega)$ may be added vectorially to the locus of

$$\frac{1}{K_dG_d(j\omega)}$$

The resulting locus has a radius vector which is

$$\frac{\theta_i}{\theta_o}(j\omega)$$

and the resonance of the final system may be studied by constructing circles of radius

$$\left|\frac{\theta_i}{\theta_o}(j\omega)\right|$$

about the origin.

Time Delays in Measuring Output

In order to control a process with a servomechanism, it is necessary to measure the output quantity. The accuracy with which the measurement is made, both statically and dynamically, is in large part indicative of the success of the controller. Unfortunately many measurements are made, for one reason or another, with equipment that responds to a sudden input with a time delay. The response delay may be either a characteristic of the measuring instrument or a characteristic of some smoothing circuit that is used.

As an illustrative problem, to show how the effects of a time delay in error measurement may be treated, a system will be discussed where the direct circuit, including control amplifier and process, is characterized by the system function

$$K_dG_d(p)=\frac{K_d}{p(T_dp+1)} \tag{11}$$

Equation 11 is representative of a cascaded control amplifier sensitive only to the input to it (and not to the derivatives or integrals of that input) and a process characterized by a characteristic time, T_d. Such a system is described in Figure 7 where

$K_d = K_{\epsilon c}K_{co}/f$ and $T_d = J/f$

This much of the system is equivalent to the type I controller of reference 1 in which the ratio of the static stiffness K to the output damping, f, is K_d, or the velocity-error constant, and the ratio of the output inertia, J, to output damping, f, is T_d.

The output measuring device of the illustrative problem will be assumed to have a response indicated by the transfer function,

$$K_fG_f(p)=\frac{1}{T_fp+1} \tag{12}$$

If the measuring device is considered by itself and is subjected to a sudden input, X_i, then equation 12 indicates that the output, X_o, will have the form

$$X_o=X_i\left(1-e^{\frac{-t}{T_f}}\right) \tag{13}$$

Before equation 11 and equation 12 are inserted in equation 9 and equation 10 to study the characteristics of the servomechanism, it is desirable to make one change of variable. If T_dp is replaced by λ; the ratio T_f/T_d defined as A; and the product K_dT_d defined as B, then the problem can be generalized, so that the solution will apply to any numerical value of T_d. Equation 11 then may be rewritten as

$$K_dG_d(\lambda)=\frac{B}{\lambda(\lambda+1)} \tag{14}$$

and equation 12 may be rewritten as

$$K_fG_f(\lambda)=\frac{1}{A\lambda+1} \tag{15}$$

The servomechanism characteristics then are determined by a substitution in equation 9 and equation 10. A nondimensional frequency variable is introduced when the operator λ is replaced by ju

$$\frac{\theta_i}{\theta_o}(ju)=\frac{ju(ju+1)}{B}+\frac{1}{jAu+1} \tag{16}$$

$$\frac{\epsilon}{\theta_o}(ju)=\frac{ju(ju+1)}{B}+\frac{1}{jAu-1}-1 \tag{17}$$

The effect of change of gain factor B for a given ratio A is illustrated in Figure 4 by a graphical construction of the problem, based on the frequency loci. As illustrated, the functions

$$\frac{1}{K_dG_d(ju)}$$

and $K_fG_f(ju)$ are plotted separately and combined by vector addition. The degree of resonance for a particular adjustment is noted by observing the proximity of the proper locus of

$$\frac{\theta_i}{\theta_o}(ju)$$

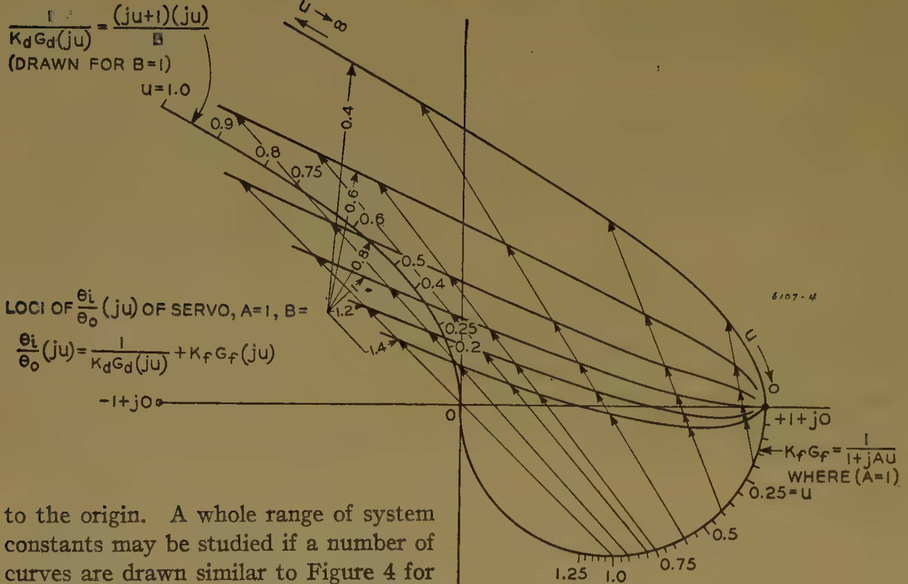

Figure 4. Loci for a proportional servomechanism with a time delay in the feed-back circuit to show the effect of change in direct circuit gain for the case of equal characteristic times in direct and feed-back circuits

to the origin. A whole range of system constants may be studied if a number of curves are drawn similar to Figure 4 for different values of A. From each of these plots a secondary group of curves may be made relating the resonant frequency and degree of resonance

$$\left(\text{minimum value of } \left|\frac{\theta_i}{\theta_o}(j\omega)\right|\right)$$

to the parameter being varied. Finally the data may be translated to the form shown in Figure 5, where lines of constant

$$\left|\frac{\theta_i}{\theta_o}(j\omega)\right|_{\text{minimum}}$$

and lines of constant resonant frequency are plotted as functions of A and B.

The process of examining the effect of change of a system parameter over a range is a laborious process. This is true to different degrees depending on the analytical technique used. Most often the analysis would not be carried so far as the one illustrated. Generally, the nature of the particular problem fixes a number of the parameters, and the problem is reduced in complexity. Quite often, too, the apparatus being considered is at hand for experimental study, and the analysis is used simply as an aid in visualizing the trend that should be observed following a change in the system.

For the problem being discussed it is practical to make a study of the transient response of the system, as well as a study of the frequency response. If the general operator λ is used instead of ju and equation 16 is combined and inverted, then

$$\frac{\theta_o(\lambda)}{\theta_i(\lambda)} = \frac{B(A\lambda+1)}{\lambda(\lambda+1)(A\lambda+1)+B} = \frac{B\left(\lambda+\frac{1}{A}\right)}{\lambda^3+\frac{A+1}{A}\lambda^2+\frac{1}{A}\lambda+\frac{B}{A}} \quad (18)$$

The denominator of the right hand side of equation 18 is the characteristic equation of the system when set equal to zero. The roots of the characteristic equation are indicative of the transient response, as has been discussed often in articles of servomechanisms.[1]

The characteristic equation for the problem being discussed is

$$\lambda^3+\frac{A+1}{A}\lambda^2+\frac{1}{A}\lambda+\frac{B}{A}=0 \quad (19)$$

which may be factored into the form

$$(\lambda+\alpha)(\lambda^2+2\zeta\mu_n\lambda+\mu_n^2)=0 \quad (20)$$

for particular values of A and B. The undamped natural frequency, μ_n, and the exponential decay constant, α, have been nondimensionalized by the use of the operator λ instead of the operator p. For a particular value of T_d the exponential characteristic time is T_d/α seconds and the undamped natural frequency of the quadratic root, ω_n, is μ_n/T_d radians per second. The damping ratio ζ, defined as the ratio of actual to critical damping of the quadratic factor, is already a nondimensional factor and does not change in value with the reversion of the problem to one of particular values for the process characteristic time T_d and measuring instrument characteristic time T_f.

If a number of solutions of equation 20 are made for several values of A when B is constant, then for the same values of A when B is constant at another value, a series of curves may be drawn which represent α, ζ, and μ_n as a function of A when B is constant. Finally the solution may be redrawn as shown in Figure 6 where lines of constant α, ζ, and μ_n are plotted as a function of A and B. This figure demonstrates immediately the results that will be obtained for a particular adjustment of parameters. There is, however, an omission, in that the magnitude of the two modes of recovery are not compared, which is not serious because it is necessary to admit at the outset, that, unless every system parameter is known

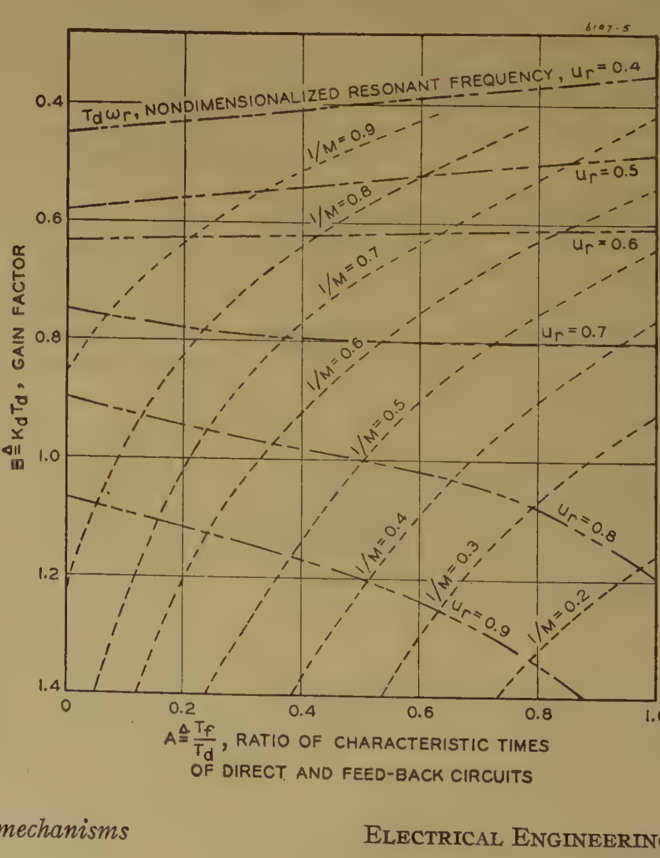

Figure 5. Resonant frequency and degree of resonance, $\left|\frac{\theta_i}{\theta_o}(j\omega)\right|_{\min} = \frac{1}{M}$, as a function of nondimensionalized system parameters of a proportional servomechanism with a time delay in the feed-back circuit

exactly and complete solutions are made for each adjustment, there will have to be a final experimental adjustment of the system. The purpose of the analysis as used in practice is to determine the approximate correct adjustment of the system and to indicate the direction implied by a proposed change of parameter values.

Because the solution to the characteristic equation of the illustrative problem can be presented fairly readily in the form of Figure 6, it is possible in this case to compare the criteria for servomechanism frequency response with criteria for transient response over a range of system adjustment by a comparison of Figures 5 and 6.

An empirical figure of merit for the design of servomechanisms by use of frequency response data has been to limit approximately the degree of resonance so that

$$M_{\max}=\left|\frac{\theta_o}{\theta_i}(j\omega)\right|_{\max}=1.5^*$$

or

$$\frac{1}{M_{\max}}=\left|\frac{\theta_i}{\theta_o}(j\omega)\right|_{\min}=0.67^{1,2}$$

For a simple system such as the one just discussed where $A=0$ and $\alpha=\infty$, there is a simple relationship between the damping ratio, ζ and the degree of resonance, $M_{\max}$. This relationship is shown graphically in Figure 7 where a value of $M_{\max}=1.5$ may be seen to correspond to a damping ratio $\zeta=0.35$. For the more complicated illustrative problem the correspondence between $M_{\max}$ and ζ can be seen to remain very similar to that for the simpler system with no delay in the feed-back circuit. This is a peculiarity of this particular problem and has been shown in unpublished work of others not to be a general case.** For more complicated systems the indications are that, when the degree of resonance is limited, the degree of stability can vary over a substantial range. A great deal still can be done to clarify the relationship between criteria for servomechanism design.

In concluding a discussion of the problem used as an illustrative example, it is desirable to point out that the characteristic equation, equation 19, is the same as any that would result from any time delay in the circuit. If, for example, the transfer function of the direct circuit were of the form

$$\frac{\theta_o}{\epsilon}=\frac{K_d}{p(T_dp+1)}\;\frac{1}{(T_fp+1)} \tag{21}$$

where the components formerly in the feed-back circuit are moved to the direct circuit, then

$$\frac{\theta_o}{\theta_i}=\frac{K_d}{p(T_dp+1)(T_fp+1)+K_d} \tag{22}$$

or

$$\frac{\theta_o(\lambda)}{\theta_i(\lambda)}=\frac{B/A}{\lambda^3+\dfrac{A+1}{A}\lambda^2+\dfrac{1}{A}\lambda+B/A} \tag{23}$$

where λ, A, and B are as defined for equation 14 and equation 15. The denominator of equation 23 is identical to that of equation 18.

The foregoing discussion leads to a general comment on the nature of the transient response resulting from any sudden disturbance, no matter where in the closed loop of a single-loop servomechanism it is inserted; namely, that the system always will recover with the same modes of oscillation and degree of stability and/or with the same exponential decay times. The predominance of one part of the response over another, however, will change, depending on where in the closed cycle the transient disturbance is introduced. This conclusion applies in a limited degree when there are multiloop systems involved.

In a similar way, the system response at the output resulting from a sinusoidal disturbance depends on where the disturb-

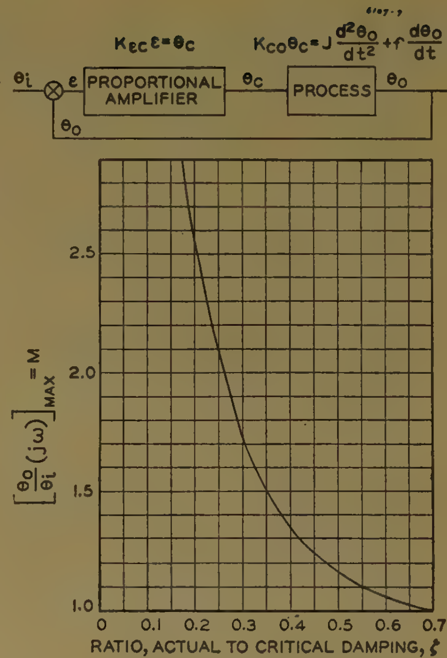

Figure 7. Correlation of degree of resonance M, with the ratio of actual to critical damping, ζ, for a proportional servomechanism

$$M=\frac{1}{2\zeta(1-\zeta^2)^{1/2}}\quad \zeta<0.707$$

$$\omega_n=\sqrt{K/J}$$

$$K=K_{ec}K_{co}$$

$$\beta=\frac{\omega}{\omega_n}$$

$$\zeta=\frac{f}{2\sqrt{KJ}}$$

Peak of frequency response occurs when

$$\beta=\sqrt{1-2\zeta^2}$$

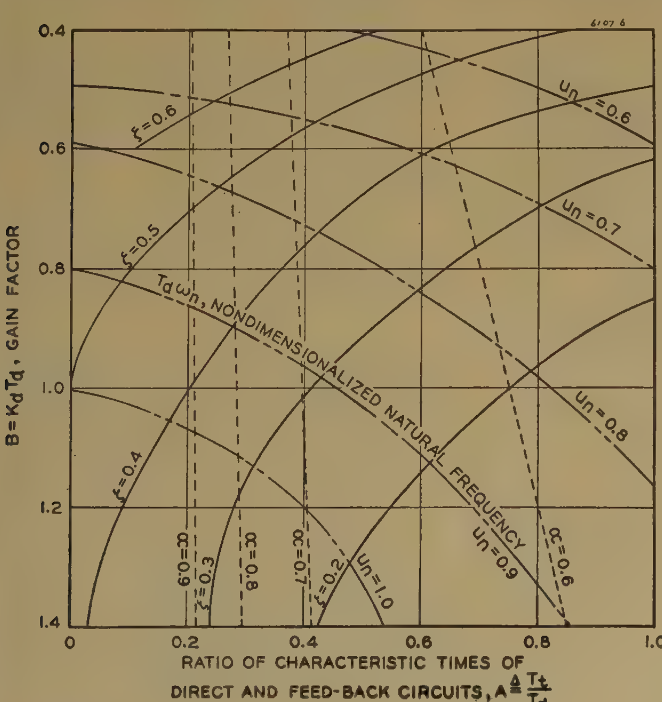

Figure 6. Presentation of the salient features of the transient response of a proportional servomechanism with a time delay in the feed-back circuit as a function of nondimensionalized system parameters

System characteristic equation

$$\lambda^3+\frac{A+1}{A}\lambda^2+\frac{1}{A}\lambda+\frac{B}{A}=(\lambda+\alpha)\times(\lambda^2+2\zeta\mu_n\lambda+\mu^2_n)$$

*The number varies considerably depending on the designer and the application.

**The manner of presentation of the roots of the characteristic equation as functions of system parameters and the comparison of these with similar data from frequency response studies is identical to that used by D. F. Tuttle of Massachusetts Institute of Technology. He compared the criteria for an integral control system.

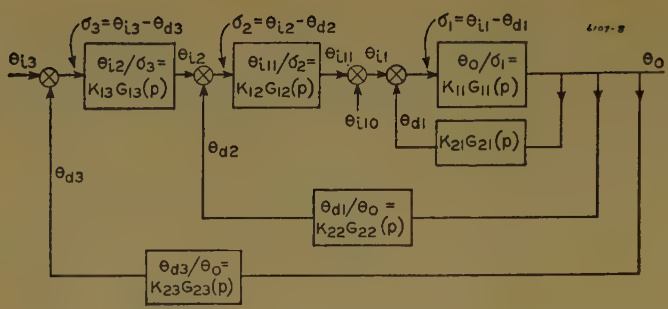

Figure 8. Diagrammatic arrangement of components in a multiloop servo system

ance is inserted. Any one system may have to be capable of satisfactory behavior following any of several disturbances, and there may therefore be a dual specification. The same general techniques of analysis employed in solving the illustrative problem can be used to determine response to a sinusoidal disturbance somewhere in the closed-cycle and hence be indicative of the transient response to a sudden disturbance at the same place. With reference to Figure 1, if the disturbance is $\theta_t(p)$ between the preamplifier and the servomotor then

$$\frac{\theta_t}{\theta_o}=\frac{1}{K_3G_3(p)K_4G_4(p)}+K_1G_1(p)K_2G_2(p) \qquad (24)$$

General Treatment of Multiloop Systems

The technique described for the study of a servomechanism whose output quantity is not measured instantaneously is a powerful tool in dealing with much more complicated systems. It is particularly useful for studying systems where there are feed-back paths within feed-back paths, or in another way of speaking, where there are several loops.

A block diagram is drawn as Figure 8 to illustrate a multiloop system in general terms. The number of loops have been limited because the treatment of a larger number of closed loops is a logical extension of the treatment of a limited number.

The reasons for a large number of closed loops in a servomechanism system may be several. There are, however, three most likely reasons for such complicated structures:

1. A servomechanism may be required to serve in more than one capacity wherein it may be used by itself or it may be used, after a switching operation, as a part of an inclusive control system.

2. It may be possible to improve servomechanism performance by the use of compensating networks whereby additional control signals are introduced, in addition to the master input, which is a function of the output quantity or its derivatives or integrals.

3. It may be necessary to measure the output quantity with a device which cannot be considered perfect as compared with the rest of the system.

The following analytical representation of the system of Figure 8 is general, in that it does not define the purpose of the several components.

(*a*). For the inside loop with others disconnected where θ_{i1} is the independent variable

$$\frac{\theta_{i1}}{\theta_o}=\frac{1}{K_{11}G_{11}(p)}+K_{21}G_{21}(p) \qquad (25)$$

(*b*). For the second loop with the outside loop disconnected, where θ_{i10} is zero and θ_{i2} is the independent variable

$$\frac{\theta_{i2}}{\theta_o}=\frac{1}{K_{12}G_{12}(p)}\left[\frac{1}{K_{11}G_{11}(p)}+K_{21}G_{21}(p)\right]+K_{22}G_{22}(p) \qquad (26)$$

(*c*). For the third loop where $\theta_{i10}=0$ and θ_{i3} is the independent variable

$$\frac{\theta_{i3}}{\theta_o}=\frac{1}{K_{13}G_{13}(p)}\left[\frac{1}{K_{12}G_{12}(p)}\left(\frac{1}{K_{11}G_{11}(p)}+K_{21}G_{21}(p)\right)+K_{22}G_{22}(p)\right]+K_{23}G_{23}(p) \qquad (27)$$

which process can be continued to include as many closed loops as there are in the system. The final result is indicative of the characteristics of the whole system.

To consider the characteristics of the inside closed loop, the outside loops may be treated as additional feed-back paths. If an input θ_{i10} is considered as adding to the output from $K_{12}G_{12}(p)$.

(*a*). For the outside loop disconnected and θ_{12} held at zero the input to the circuit $K_{11}G_{11}(p)$ is

$$\theta_{i10}+[K_{12}G_{12}(p)K_{22}G_{22}(p)]\theta_o-[K_{21}G_{21}(p)]\theta_o \qquad (28)$$

and this is reproduced at the output as

$$K_{11}G_{11}(p)[\theta_{i10}+K_{12}G_{12}(p)K_{22}G_{22}(p)\theta_o-K_{21}G_{21}(p)\theta_o]=\theta_o \qquad (29)$$

or

$$\frac{\theta_{i10}}{\theta_o}=\frac{1}{K_{11}G_{11}(p)}+K_{21}G_{21}(p)-K_{12}G_{12}(p)K_{22}G_{22}(p) \qquad (30)$$

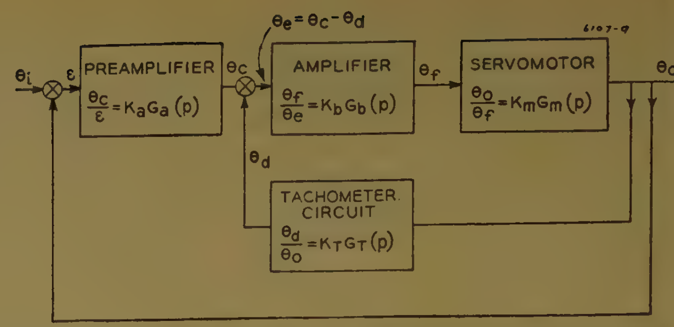

Figure 9. Diagrammatic arrangement of components for use of tachometric feedback in a positional servomechanism

which is a result which can be obtained by the addition of a new vector locus to the original one, equation 25, for the inside loop with the others disconnected.

(*b*). If the third loop as well as the second loop is connected and θ_{13} held at zero, then the quantity σ_2 is the difference between θ_{i2} and θ_{d2}, and this is a simple addition to equation 30

$$\frac{\theta_{i10}}{\theta_o}=\frac{1}{K_{11}G_{11}(p)}+K_{21}G_{21}(p)+K_{12}G_{12}(p)\times[K_{23}G_{23}(p)K_{13}G_{13}(p)-K_{22}G_{22}(p)] \qquad (31)$$

The foregoing process may be continued to include as many loops as are necessary. Even though the systems are very complicated, there is a reasonable way to consider the effect of change in any of the circuits by graphical construction of the loci of the reciprocal system function, where the operator is ($j\omega$), and by observation of its salient characteristics.

Comments on Compensating Networks

The performance required of a given servomechanism is derived from the specifications for the operation of the whole control system. Such specifications are particular to any one application but are generally of two types

1. Those concerned with the steady state response to a typical input.

2. Those concerned with the dynamic or transient response.

In the first group, familiar specifications are concerned with: static accuracy; steady-state following error when the input is uniformly changing; steady-state sinusoidal errors when the input is moving sinusoidally at low frequencies; errors introduced by the application of load on the output. The second group of specifications is concerned primarily with the speed of response and degree of stability of the system, following suddenly applied inputs or loads. The problem for the designer of servomechanisms usually re-

solves itself into determining means for obtaining the desired steady-state response without harming the dynamic response characteristics.

In accomplishing this purpose in designing servomechanisms, the engineer usually resorts to the use of certain general techniques which are economic of equipment and which often make the control system feasible. These techniques center around

1. The close specification of the characteristics of the process, so that it is more controllable.

2. The use of special control circuits.

One widely used technique is to make the control amplifier sensitive to derivatives and integrals of the servo error as well as the error itself. The controller then is described as one which anticipates and remembers. The general purpose of the use of derivatives of error is to improve the degree of stability of the system, thereby making possible larger sensitivity. The general purpose in the use of integrals of the error is to increase the system sensitivity statically or at low frequencies without materially harming the performance of the system at frequencies where it is likely to oscillate. The techniques for the design of these compensating devices are not the main concern of this paper. They have been discussed widely (see reference list) and can be treated as cascaded elements in the system. In Figure 1, a change in the controller, represented by the operational transfer function $K_1G_1(p)$, can be treated readily since it is multiplied by the transfer functions of the remaining cascaded elements. This has been indicated previously in equation 1.

There are a number of other compensating networks which cannot be accurately described as cascade elements in the control system. Several examples of these are enumerated:

1. *Feedback around a component or group of components.* This often is done to improve the dynamic response of elements or to remove the dependence on open-loop instrumentation in the action of an element. Integral controllers sometimes are made by the use of feedback.

2. *Output feedback.* It is often possible to obtain a stabilizing device by returning signals to the controller which are functions of the derivatives of the output quantity, as well as functions of the output quantity itself. When these signals are used without subtracting them from similar functions of the input, then they are no longer cascade elements.[1,8]

3. *Input feed-forward.* This is a term used to describe systems which measure a function of the derivatives of the input as well as the input itself, thus supplying the control system with advance knowledge of what is going to be required of it.[8]

4. *Tuned loads.* These are used to impose a restraint on the output member which is frequency sensitive. They are devices that limit the ability of the output member to oscillate at the resonant frequency of the system.

All of the afore-mentioned techniques are fairly complicated in their application. It is not practicable to treat them all, and the opportunity to obtain general solutions is limited. They can, however, be studied quite satisfactorily by the use of appropriate system functions. The remainder of this paper will deal with an illustrative example wherein output feedback is used to improve the performance of the system.

Tachometric Feedback in Positional Servomechanisms

If the speed of the output of a position control is measured, and a function of this signal is fed back degeneratively into the amplifier stages, where it is added to the positional error signal, the performance of the servomechanism may be improved. This may be visualized physically by comparing this technique with that of employing the derivative of the servo error. When the servo input is stationary, the two are identical in function if the tachometric signal is not modified by some additional network. It would be concluded, therefore, that tachometric feedback has a stabilizing effect, since the good use of error derivative techniques is well known.[1,2]

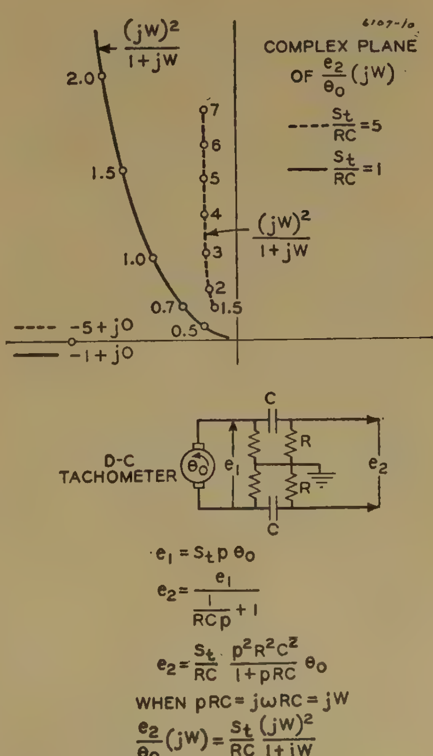

Figure 10. D-c tachometer with high-pass filter

Actually the tachometer signal usually is modified to improve the steady-state error that results when the input is moving at a constant velocity. A high-pass filter usually is used, so that no signal is sent back when the output is moving at a constant velocity or is following sinusoidal inputs of relatively low frequencies. The filter, however, allows the tachometer signal to pass freely at frequencies where the servomechanism is likely to oscillate. A simple filter is shown in a tachometer circuit as a part of Figure 10. By this technique the stabilizing effect of the tachometer is made available at frequencies where the servomechanism is likely to oscillate and does not require an even greater error signal to make the output move at a given speed.

The tachometer signal need not be added to the control system at the same place where the positional error signal is introduced. A general arrangement of components is illustrated in block diagram form in Figure 9. The preamplifier may use an integrating circuit or some other compensating circuit if the specifications for the servomechanism cannot be met without their use. For the first example to be discussed here the amplifier will be assumed to be proportional and characterized by a gain factor, K_a. Also in the first example it will be assumed that the servomotor is one in which the output damping is negligible and that the inertia of the motor and output member is the only energy storage element on the output shaft. Under these conditions the transfer function $K_mG_m(p)=K_m/J_mp^2$. Without the use of some sort of compensating device this system would be unstable. To stabilize the system let the tachometer circuit of Figure 10 be used. The amplifier $K_bG_b(p)$ will be assumed to be proportional and to have no time lags. Equations for the system may be written on the basis of the foregoing simplifications:

$$\theta_e=\theta_c-\theta_d=K_a\epsilon-\frac{S_tRCp^2\theta_o}{RCp+1} \tag{32}$$

$$\theta_e=\frac{J_mp^2}{K_bK_m}\theta_o \tag{33}$$

from which

$$\frac{\epsilon}{\theta_o}=\frac{1}{K_a}\left[\frac{J_mp^2}{K_bK_m}+\frac{S_tRCp^2}{RCp+1}\right] \tag{34}$$

and since by definition, $\epsilon=\theta_i-\theta_o$

$$\frac{\theta_i}{\theta_o}=1+\frac{1}{K_a}\left[\frac{J_mp^2}{K_bK_m}+\frac{S_tRCp^2}{RC_p+1}\right] \tag{35}$$

which may be written in the form

$$\frac{\theta_i}{\theta_o}=1+\frac{S_t}{K_aRC}\left[\frac{J_m}{RCK_bK_mS_t}\lambda^2+\frac{\lambda^2}{\lambda+1}\right] \quad (36)$$

where $\lambda=RCp$. If the right hand side of equation 35 is expressed in terms of a common denominator, the numerator of the resulting fraction is the characteristic equation when equated to zero.

$$J_mRCp^3+(K_bK_mS_tRC+J_m)p^2+K_aK_bK_mRCp+K_aK_bK_m=0 \quad (37)$$

The characteristic equation may be written in terms of λ instead of p and two nondimensional parameters defined

$$F=\frac{J_m}{RCK_bK_mS_t} \quad (38)$$

$$D=\frac{S_t}{K_aRC} \quad (39)$$

$$\lambda^3+\left(1+\frac{1}{F}\right)\lambda^2+\frac{1}{FD}\lambda+\frac{1}{FD}=0 \quad (40)$$

Equation 40 is in a form where it may be broken down readily into a study of the real and quadratic factor. These factors could be expressed graphically as functions of F and D in a manner similar to Figure 6.

An important deduction can be made with reference to equations 38 and 39. It can be seen that F and D may be controlled separately in magnitude by adjustment of the time constants and component sensitivities in the system. This means that mathematically any desired result may be obtained for the response of the system.

There are two practical limitations on the procedure of unlimited adjustment. The first, and most important, is that no physical system can be represented by such simple differential equations as a first order approximation when the natural and resonant frequencies for the system become very high. The second limitation is that the specifications for the performance of the servomechanism do not, as a rule, justify the expense in equipment in obtaining better performance than is required.

By using the same substitutions given in equations 37, 38, and 39 the frequency response loci may be studied in accordance with the functional relationship indicated by equation 35. Setting $\lambda=jW$

$$\frac{\theta_i}{\theta_o}(jW)=1+D\left[F(jW)^2+\frac{(jW)^2}{1+jW}\right] \quad (41)$$

Figure 11 shows an example of the construction of typical loci for one particular value of D and two values of F. In drawing these loci, it would have been as easy to construct them by adding to the locus of

$$\frac{(jW)^2}{1+jW}$$

shown in Figure 10, the locus of $F(jW)^2$ and work with the resultant locus as a reciprocal system function

$$\frac{\epsilon}{\theta_o}(jW)$$

The polar expansion of this locus is controlled by the gain factor D.

The same conclusion about the adjustment of the system may be reached from a study of the frequency loci as was reached from a study of the characteristic equation; namely, that any response desired could be obtained by appropriate selection of parameter values.

There is, however, one important failing of this technique in that there is no simple way known to the author that has yet been evolved to see the presence of predominating real roots in the transient response by simple examination of the locus. It has been suggested that the rate of change of phase near zero frequency might hold a clue, but this has not been investigated thoroughly. Even in a complicated system, however, it is not a long problem to evaluate a real root and its coefficient for particular values and for a specified disturbance. Hence, there is the desirability of always keeping in close touch with the transient specifications of a system when the major part of the study is done by frequency characteristics.

It will be desirable to conclude this discussion of tachometer feedback by studying the application to a system more exactly represented by higher order differential equations than the afore-mentioned one. In doing this, only the frequency response technique will be studied, because a study of the transient response becomes hopelessly complicated, except for final checks on the presence of undesirable real roots in the characteristic equation which are excited by typical inputs expected in the application.

A typical servomotor for a position controller will be an integrating device with an elastance storage element as well as an inertia storage element. The transfer function, K_mG_m, for such a servomotor would be of the form

$$K_mG_m(p)=\frac{S_m}{p(J_mp^2+f_mp+k_m)} \quad (42)$$

where K_m is S_m/k_m and is expressible as per unit speed output per unit input. If the numerator and denominator of equation 44 are divided by k_m, and the open-loop undamped natural frequency of this component be written as $\omega_n=k_m/J_m$, and the ratio of actual to critical damping of the open-loop characteristic be written as

$$\zeta=\frac{f_m}{2\sqrt{J_mk_m}}$$

then equation 44 may be rewritten

$$K_mG_m(p)=\frac{\dfrac{K_m}{\omega_n}}{\dfrac{1}{\omega_n}p\left(\dfrac{1}{\omega_n^2}p^2+\dfrac{2\zeta}{\omega_n}p+1\right)} \quad (43)$$

and transformed, when

$\lambda=\dfrac{1}{\omega_n}p$ into

$$K_mG_m(\lambda)=\frac{S_m/k_m\omega_n}{\lambda(\lambda^2+2\zeta\lambda+1)} \quad (44)$$

The amplifier used with such a power motor probably has an open-loop speed of response of the same order of magnitude as the servomotor itself. If the transfer function of the amplifier be assumed to be of the form

$$K_bG_b(p)=\frac{K_b}{\tau p+1} \quad (45)$$

where the characteristic time, τ, is arbitrarily set equal to $1/\omega_n$, which is commensurate with the foregoing statement regarding its speed of response, equation 45 then may be rewritten in terms of

$$K_bG_b(\lambda)=\frac{K_b}{\lambda+1} \quad (46)$$

The cascade of the amplier and servomotor then may be written as

$$K_mG_m(p)K_bG_b(p)=\frac{\dfrac{K_bK_m}{\omega_n}}{(\lambda+1)(\lambda^2+2\zeta\lambda+1)\lambda} \quad (47)$$

The preamplifier will be assumed to be characterized by a gain K_a and negligible characteristic time.

If the components are assembled in a servomechanism without tachometer feedback and the gain of the system

$$\frac{K_aK_bK_m}{\omega_n}$$

be adjusted to give the highest resonant frequency possible with a peak degree of resonance

$$\left|\frac{\theta_i}{\theta_o}\right|_{\min}$$

of 0.7, then the resulting frequency locus will be as shown in Figure 12 where $\zeta=0.3$ and $\lambda=j\mu$. The gain adjustment was such as to make

$$\frac{K_aK_bK_m}{\omega_n}=\frac{1}{2.8}$$

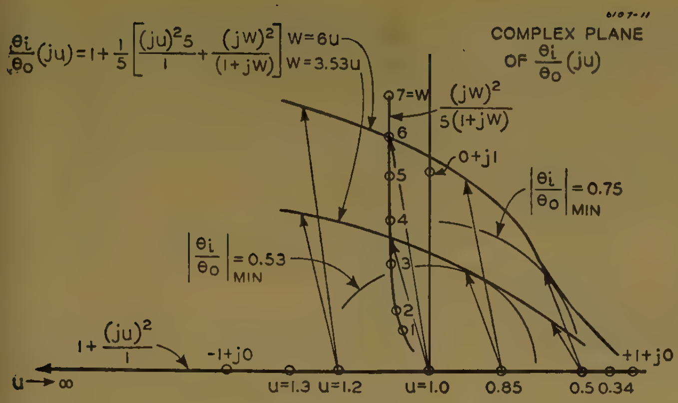

Figure 11. Loci illustrating a study of the application of tachometer feedback to a proportional servomechanism whose output member is characterized by load inertia and no damping

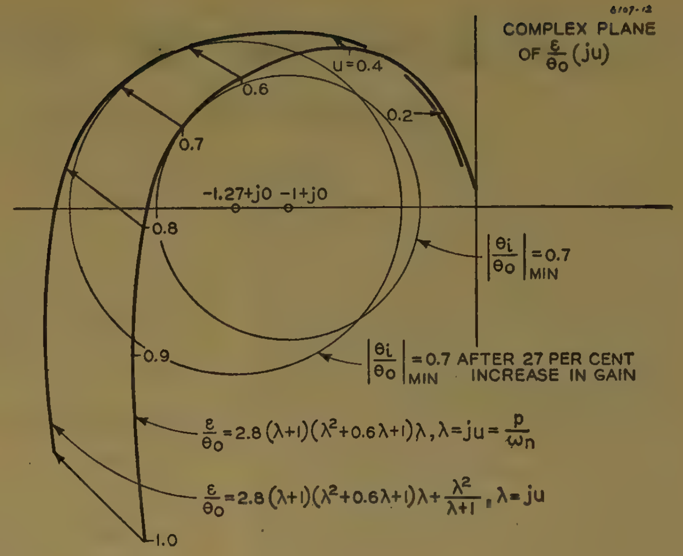

Figure 12. Loci illustrating a study of the application of tachometer feedback to a positional servomechanism with four energy storage elements

If, now, tachometer feedback is to be added to the system, the analytical combination of the system is indicated in equation 48.

$$\frac{\epsilon}{\theta_o}=\frac{1}{K_a}\frac{1}{K_mG_m(p)K_bG_b(p)}+K_tG_t(p) \qquad (48)$$

which is derived for the system block diagram of Figure 9.

The type of tachometer feedback used will be that illustrated in Figure 10, so that

$$K_tG_t(p)=-\frac{\frac{S_t}{RC}R^2C^2p^2}{RCp+1} \qquad (49)$$

A question immediately arises about the magnitude of the tachometer signal that can be made available, or in other words what is a reasonable value of K_t with respect to the sensitivity of the amplifier and servomotor in cascade. If K_t were exactly equal to $1/K_mK_b$ then, in the absence of the tachometer RC network, the amplifier K_a would have to supply just twice as much signal to make the output run at a given speed as it would in the absence of the tachometer. Calculation would show that a locus resulting from substitution in equation 50 would be altered very little by the addition of tachometer feedback with so low a tachometer sensitivity, unless the characteristic time of the RC network is made very large. Unfortunately, if the characteristic time of the RC network is made much larger than those in the remainder of the system, the speed of response of the servomechanism will be reduced by the introduction of a predominant exponential decay in the transient response. This conclusion is largely an experience deduction based on heuristic reasoning about the flow of correcting signals through the circuits. However, it has been substantiated by actual calculation in a number of instances.

In Figure 12 the original locus is shown modified by a tachometer feed-back circuit with the particular adjustment

$$RC=\frac{1}{\omega_n} \qquad (50)$$

and

$$K_t=\frac{S_t}{RC}=\frac{\omega_n}{K_aK_mK_b}=2.8 \qquad (51)$$

The addition of tachometer feedback allows an increase in the preamplifier gain of 27 per cent without increasing the degree of resonance.

The allowable increase in gain may be obtained by evaluating the polar contraction of the final locus before again becoming tangent to the $1/M=0.7$ circle. Another technique is to find by trial and error a new circle, whose radius is 0.7 times the length along the axis to the center, that will be tangent to the new locus. The amount by which the gain may be increased is read immediately from the scale on the negative real axis.

In summary it cannot be emphasized too strongly that the major value of such analytical techniques as are described here is to give a quantitatively approximate idea of the correct approach to a problem. However, when the analysis includes experimental data on the performance of the major components, then the design of the remaining circuits can be done very exactly, although always subject to check on the applicability of the criteria used for design.

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A Fractional Termination for Ladder Networks

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Synopsis: When using a uniform ladder network to simulate a smooth transmission line, the possibility is considered of deriving a termination which would be intermediate between the usual mid-shunt and mid-series types and which would have advantages over the latter. It is found that a new termination is possible which gives some improvement throughout a limited frequency range. For the dissipationless case, the range corresponds to a line of less than a quarter wave length. The improvements attainable are of the order of a few per cent and become less significant the greater the number of network sections. Thus, although the formula for the new termination and the estimates of the results which it will accomplish are given for any number of sections, the new termination is of most practical value for circuits of few sections. In arriving at this termination, consideration is given to the variation of each of the elements of the pair on each end; but of these two, the end element is the only one whose variation yields an improvement. For the frequency range for which the modified circuit is an improvement, the distinction between the two original types of termination largely is lost, the modified midshunt and modified mid-series terminations being almost identical. Beyond this range there is a difference between them, and the corresponding original and new networks are nearly equivalent. A possible application of the results is suggested in which the simulation of a smooth line is not involved.

IT is well known that by taking a sufficient number of sections, a mid-shunt or mid-series terminated uniform ladder network may be made to approximate a smooth line as closely as desired over a low-pass band of frequencies the width of which increases as the number of network sections is increased. It is also true, considering the network in the 4-terminal sense, that under some conditions a mid-shunt termination gives the best results while with other terminal conditions the mid-series arrangement is superior. The question to be treated is whether an intermediate termination can be found which will give a uniformly good approximation for all conditions and, if possible, a superior performance. This is to be investigated with the understanding that the network must retain its 4-terminal identity, which implies symmetry about the line indicated in Figure 1. This rules out known nonuniform networks that offer economy of circuit elements when the line is to be represented in a 2-terminal sense only.[1]

Properties of Uniform Ladder Networks

It is considered that an analysis of the ladder network, when terminal quantities are involved, is most reasonably carried out in terms of the 4-terminal network parameters A, B, and C, which are defined by the equations

$$E_S = AE_R + Z_T BI_R \tag{1a}$$

$$I_S = Y_T CE_R + AI_R \tag{1b}$$

in the symmetrical case. The voltages and currents involved are indicated in

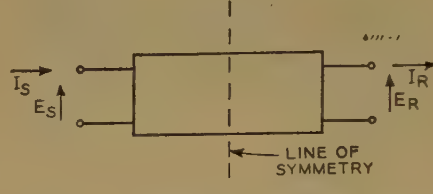

Figure 1. Notation for a 4-terminal network

Figure 1, and Z_T and Y_T are respectively the total series impedance and shunt admittance of the line. This notation is slightly different from the usual practice in that B and C are written respectively with the factors Z_T and Y_T removed. This change simplifies many of the subsequent equations. One may obtain the quantities that are necessary for expressing the circuit in other forms from equations involving A, B, and C, which may be found in standard texts on circuit analysis.[2]

Expressions for A, B, and C in power series of the variable Z_TY_T have been given in a previous paper[3] for a uniform ladder network with either type of termination. The series form is used because it affords simple comparisons with corresponding formulas for a smooth line. These formulas are repeated here for easy reference, but with primes on all quantities so that unprimed notation may be used for the derived circuits. For a network of M identical T sections connected in cascade we define

$$\alpha' = \sqrt{Z'_T Y'_T} \tag{2}$$

and have

$$A' = \sum_{\lambda=0}^{M} a'_\lambda (\alpha')^{2\lambda} \tag{3a}$$

$$B' = \sum_{\lambda=0}^{M} b'_\lambda (\alpha')^{2\lambda} \tag{3b}$$

$$C' = \sum_{\lambda=0}^{M-1} c'_\lambda (\alpha')^{2\lambda} \tag{3c}$$

in which

$$a'_\lambda = \frac{1}{(2\lambda)!} \prod_{n=0}^{\lambda-1} \left(1 - \frac{n^2}{M^2}\right) \tag{4a}$$

$$b'_\lambda = \frac{1}{(2\lambda+1)!} \left(1 + \frac{\lambda}{2M^2}\right) \prod_{n=0}^{\lambda-1} \left(1 - \frac{n^2}{M^2}\right) \tag{4b}$$

$$c'_\lambda = \frac{1}{(2\lambda+1)!} \prod_{n=0}^{\lambda} \left(1 - \frac{n^2}{M^2}\right) \tag{4c}$$

The corresponding results for M identical π sections in cascade are obtained by interchanging B' and C',[3] and the corresponding coefficients for a smooth line are obtained from equations 3 and 4 by allowing M to become infinite. The parameters and coefficients for the smooth line subsequently will be referred to as ideal values and will be written with a bar above the appropriate symbol.

Modified Terminations

The simple relationship between the B and C parameters of the T and π section constructions maintains even when all network sections are not identical, so long as the modifications of corresponding elements of the two networks are related dually.[3] That is, so long as a given change in a series element of one network is represented by an equivalent change in the corresponding shunt element of the other network, we need treat only one network. This enables us to proceed by analyzing only the T-section case.

The algebra is simplified if we assume an ordinary mid-series terminated uniform network on each end of which is connected a new series element. The new total series impedance will be designated by Z_T, and the element added to each end will be of magnitude ρZ_T. These definitions are illustrated in Figure 2A, along with Figure 2B, which shows

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the equivalent π-section case, the results for which will be included, as explained above.

For the new circuit

$$\alpha=\sqrt{Z_T Y_T} \tag{5a}$$

$$Z'_T=(1-2\rho)Z_T \tag{5b}$$

$$Y'_T=Y_T \tag{5c}$$

$$\alpha'=\alpha\sqrt{1-2\rho} \tag{5d}$$

No adjustment is possible if the network has only a single section, so the number of sections (M) must be greater than or equal to 2. The formulas for A, B, and C, the new network parameters, are easily obtained by considering the network to be formed by the cascade connection of three elemental networks as indicated in Figure 2. This connection lends itself to the use of matrix algebra in the following way:[4]

$$\begin{pmatrix} A & BZ_T \\ CY_T & A \end{pmatrix}=\begin{pmatrix} 1 & \rho Z_T \\ 0 & 1 \end{pmatrix}\times \begin{pmatrix} A' & B'Z'_T \\ C'Y'_T & A' \end{pmatrix}\begin{pmatrix} 1 & \rho Z_T \\ 0 & 1 \end{pmatrix}= \begin{pmatrix} A'+\rho C'\alpha^2 & [2A'\rho+B'(1-2\rho)+\rho^2C'\alpha^2]Z_T \\ C'Y_T & A'+\rho C'\alpha^2 \end{pmatrix} \tag{6}$$

The formulas for A, B, and C may be picked from the matrix on the right of the above equation. Equations 3a–3c are used for A', B', and C' with the observation that the $(\alpha')^{2\lambda}$ factors may be replaced by $\alpha^{2\lambda}(1-2\rho)^{\lambda}$. Also, noting that whenever the term α^2 appears in the final matrix of equation 6, it always carries C' as a coefficient, and that the maximum exponent in the series for C' is two less than in the series for A' or B', it follows that the maximum exponents of α in the series for A, B, and C will be the same as in equations 3a–3c. Hence

$$A=\sum_{\lambda=0}^{M} a_\lambda\alpha^{2\lambda} \tag{7a}$$

$$B=\sum_{\lambda=0}^{M} b_\lambda\alpha^{2\lambda} \tag{7b}$$

$$C=\sum_{\lambda=0}^{M-1} c_\lambda\alpha^{2\lambda} \tag{7c}$$

Formulas for a_λ, b_λ, and c_λ are obtained from the individual elements of the last matrix of equation 6, in conjunction with equations 3a–3c. The results are

$$a_\lambda=a'_\lambda(1-2\rho)^\lambda+\rho c'_{\lambda-1}(1-2\rho)^{\lambda-1} \tag{8a}$$

$$b_\lambda=2\rho a'_\lambda(1-2\rho)^\lambda+b'_\lambda(1-2\rho)^{\lambda+1}+\rho^2c'_{\lambda-1}(1-2\rho)^{\lambda-1} \tag{8b}$$

$$c_\lambda=c'_\lambda(1-2\rho)^\lambda \tag{8c}$$

In formula 8a, a negative subscript is understood to indicate a value of zero for that coefficient. At this point it is well

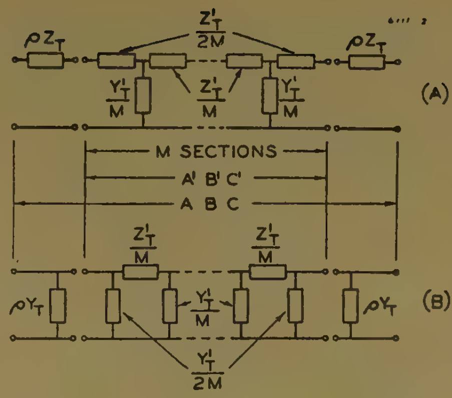

Figure 2. Single-element modification

to make the cautionary remark that if these are to be interpreted for the shunt-terminated case, it is only the formulas for b_λ and c_λ that should be interchanged. Reflection will show that b'_λ and c'_λ must not be interchanged if they are still given by equations 4a–4c. Of course, if an exchange is made there also, the change should be made in equations 8a–8c as well. In other words, if b'_λ and c'_λ are not interchanged, they retain their significance of respective coefficients of B' and C' for a uniform mid-series terminated network, while if they are interchanged, their new significance relates to a mid-shunt terminated uniform network.

The condition for the new termination has been stated as the removal of the distinction between series and shunt element termination with, if possible, an improvement in the approximation for a smooth line. The non-equivalence of the two ordinary terminations arises from the non-equality of b'_λ and c'_λ in equations 4a–4c. One therefore proceeds by first making $b_\lambda=c_\lambda$ for as many values of λ as possible, starting from $\lambda=0$, and later checking whether they are less in error than their corresponding primed coefficients. From equations 4a–4c it is seen that a'_0, b'_0, and c'_0 each has its correct value, and from equations 8a–8c it follows that

$$a_0=b_0=c_0=1 \tag{9}$$

It also is seen from equation 8a, and the fact that $a'_1=1/2$ and $c'_0=1$, that

$$a_1=a'_1(1-2\rho)+\rho c'_0={}^1/_2 \tag{10}$$

Thus, to have the first terms correct in the power series expansions of B and C and to have the first two terms correct in the expansion of A, it is only necessary to have the total series impedance and shunt admittance equivalent to those of the simulated line so α will have the correct value.

Attention logically is turned next to a_2, b_1, and c_1. It is shown in Appendix I that these are interrelated by the formula

$$a_2=\frac{b_1+c_1}{2}-\frac{1}{8} \tag{11}$$

so a_2 will be automatically determined when b_1 and c_1 are established. We shall proceed by putting $\lambda=1$ in equations 8b, 8c, and so on to obtain, after substituting values for the primed coefficients from equations 4a–4c, the following:

$$b_1=\rho(1-2\rho)+(1-2\rho)^2\left(\frac{1}{6}\right)\left(1+\frac{1}{2M^2}\right) \tag{12a}$$

$$c_1=(1-2\rho)\left(\frac{1}{6}\right)\left(1-\frac{1}{M^2}\right) \tag{12b}$$

Formulas 12a and 12b may be equated to yield a quadratic equation in ρ the solution of which readily is found to be

$$\rho=1-\frac{1}{2}\sqrt{\frac{4M^2-1}{M^2-1}} \tag{13}$$

The negative sign is chosen for the radical from knowledge that $\rho=0$ must correspond to infinite M. It is interesting to note that ρ is negative, implying a reduction of the end element. This is reasonable because it physically means a compromise between the two ordinary terminations. The substitution of this value of ρ in equation 12b gives

$$b_1=c_1=\frac{1}{6}\sqrt{1-\frac{1}{M^2}}\times\left(2\sqrt{1-\frac{1}{4M^2}}-\sqrt{1-\frac{1}{M^2}}\right) \tag{14}$$

It remains to be seen whether this value is less in error than b'_1 and c'_1. It should be noted parenthetically that when comparisons are to be made between the new and the old coefficients, the latter must represent a network having the same overall impedance and admittance as the new circuit. This only affects A', B', and C', through Z'_T, Y'_T, and α', and not a'_λ, b'_λ, and c'_λ. It is only the interpretation which is affected by varying Z'_T, and Y'_T. This is equivalent to saying that the primes may be removed from Z'_T, Y'_T, and α', but that they must be retained on the parameters and coefficients. It will be assumed that this is done for all the subsequent comparisons.

The error of $b_1=c_1$, as given by equation 14, may be estimated by the process shown in Appendix II. It is found that

$$\frac{1}{24M^2}<\frac{1}{6}-c_1=\frac{1}{6}-b_1<\frac{3}{49M^2} \tag{15}$$

while the errors of b'_1 and c'_1 are, from equations 4a–4c,

$$b'_1-\frac{1}{6}=\frac{1}{12M^2} \tag{16a}$$

$$\frac{1}{6}-c'_1=\frac{1}{6M^2} \tag{16b}$$

It thus is assured that the errors in b_1 and c_1 are less than those in b'_1 and c'_1.

An estimate of the error in the corresponding a_2 may be obtained from equation 11 and inequality 15. One directly obtains

$$\frac{1}{24M^2}<\frac{1}{24}-a_2<\frac{3}{49M^2} \tag{17}$$

It is also important to know how it compares with its value when $\rho=0$. An estimate of the ratio a_2/a'_2 can be found by a procedure shown in Appendix III which gives

$$\frac{(M^2-1/4)(M^2-7/4)}{(M^2-1)^2}<\frac{a_2}{a'_2}<\frac{(M^2-2/5)(M^2-8/5)}{(M^2-1)^2}<1 \tag{18}$$

This remains near, but less than, unity. The estimate is less accurate when M is small, so actual values of the ratio are tabulated in Table I along with the ratios of a'_2, a_2, b'_1, $b_1=c_1$, and c'_1 to their ideal values. One must observe that the error in a_2 is greater than the error in a'_2. This follows from equation 4a, which shows that a'_2 is less than $\bar{a}_2$, and from relation 18, which shows a_2 to be less than a'_2.

When $|\alpha|$ is sufficiently small, only the first two terms in each of the series for A', B', and C' will be important because it is known that the coefficients decrease with increasing λ. Therefore if the coefficients for the new circuit, when λ is greater than or equal to 2, are not brought into prominence by the modification, the new circuit will have less error and will be more symmetrical in B and C, for this range of $|\alpha|$. Estimates of a_λ, b_λ, and c_λ are therefore in order for $\lambda \geqq 2$. The methods used in obtaining these estimates are illustrated in Appendix III, with the results given in the following inequalities:

$$\left(\frac{M^2-2/5}{M^2-1}\right)^{\lambda-1}\left(\frac{M+1/4}{M+1}\right)<\frac{a_\lambda}{a'_\lambda}<\left(\frac{M^2-1/4}{M^2-1}\right)^{\lambda-1}\left(\frac{M^2-11/5}{M^2-1}\right) \qquad \lambda\geqq 3,\ M\geqq 3 \tag{19a}$$

$$\left(\frac{M^2-2/5}{M^2-1}\right)^{\lambda-1}\left(\frac{M-1}{M+1}\right)<\frac{b_\lambda}{b'_\lambda}<\left(\frac{M^2-1/4}{M^2-1}\right)^{\lambda-1} \qquad \lambda\geqq 2,\ M\geqq 4 \tag{19b}$$

$$\left(\frac{M^2-2/5}{M^2-1}\right)^{\lambda}<\frac{c_\lambda}{c'_\lambda}<\left(\frac{M^2-1/4}{M^2-1}\right)^{\lambda} \qquad \lambda\geqq 2,\ M\geqq 2 \tag{19c}$$

The second of these is valid only for $M\geqq 4$ for a reason which is pointed out with its derivation. The actual values of b_λ/b'_λ for the range not covered by relation 19b have been computed and are as follows: with $M=2$ and $\lambda=2$, ratio$=0.7221$; with $M=3$ and $\lambda=2$ and 3, ratios equal 0.8505 and 0.7965, respectively. A statement about the application of these estimates for the shunt terminated case is necessary to avoid confusion. In the light of the statement following equations 8a–8c, one would correctly conclude that b'_λ and c'_λ retain their positions, while b_λ and c_λ are interchanged. However, this leaves the ratios in unreasonable forms, so it is better here to interchange b'_λ and c'_λ with the understanding that they are given by formulas 4a–4c in an interchanged condition. This is in agreement with the alternative proposition contained in the statement referred to. Relations 19a–19c make it clear that none of the coefficients is greatly changed by the modification.

Before continuing with the investigation, the meaning of these results will be discussed rather fully because they have a bearing on the development to follow. It has been pointed out that the B and C coefficients for the new circuit are more nearly equal, and are nearer the ideal value, only so long as $|\alpha|$ is sufficiently small to make the first two terms predominant in the series for A, B, and C. It would be convenient to be able to specify this range definitely, but it depends on the circuit termination and on the angle of the complex number α. Something can be said about the range, however. In the following discussion the range of $|\alpha|$ over which the new circuit is an improvement will be referred to as the useful range. From equation 11, if we take Δb_1 and Δc_1 as the respective errors of b_1 and c_1, it follows that the error (Δa_2) in a_2 is

$$\Delta a_2=\frac{\Delta b_1+\Delta c_1}{2} \tag{20}$$

Considering these terms only, we see that when $|\alpha|=1$ the error in A caused by Δa_2 is as great as the averaged errors of B and C caused by Δb_1 and Δc_1. For larger $|\alpha|$ the error of A will become rapidly worse because a_2 is the coefficient of α^4, while b_1 and c_1 are coefficients of α^2. Thus it would seem that the useful range is about $|\alpha|<1$. However, when considering a_2, it must be remembered that its error is only slightly less in the original circuit. Since we are discussing improvements, although a_2 may cause more error in A than the errors of b_1 and c_1 cause in B and C, the new circuit actually still may be better so long as B and C are more accurate. This argument is reinforced by the realization that in use the B and C parameters have multiplying factors which depend on Z_T and Y_T, and on the terminal conditions. Thus A, B, and C may have different effects on the performance and, so long as B and C are less in error and the error in A is not too much greater, the new circuit may be better. An exceptional case would be one in which the original circuit errors in A and B or C were mutually compensating. In view of these ideas, it would seem logical to define the useful range as that range for which the errors in B and C remain less than in the original circuit, with the understanding that some cases may arise for which the new circuit is not better beyond $|\alpha|=1$. This still does not lead to a unique definition, but it does lead to the expectation of a range somewhat greater than $|\alpha|<1$ because b_2 and c_2 are each much less than a_2 and do not need to be considered until $|\alpha|$ is considerably greater than unity.

Some estimates of the useful range for the dissipationless case may be obtained from Figure 3. A', B', C', A, B, and C are shown as functions of $\alpha=j\omega\sqrt{LC}$ (L and C being the respective total inductance and capacitance of the line) for three values of M. From these graphs a range of $|\alpha|<1.5$ seems to be reasonable for this condition. This range is sensibly independent of M, a consequence of having improved only b_1 and c_1 in all cases. The only advantage of increasing M is to make the errors smaller. This statement refers of course only to the useful range. Increasing M reduces the errors of all circuits for larger $|\alpha|$.

The estimates given in relation 19 are applicable in judging the relative behavior of the circuits beyond the useful range. The percentage variations of the coefficients of order higher than $\lambda=1$ may be of the same order of magnitude as the percentage improvements realized in b_1 and c_1. However, the latter are significant because those coefficients are nearly correct, while for the former the errors are so large as to make the changes of

Table I

M	$a'_2/\bar{a}_2$	$a_2/\bar{a}_2$	a_2/a'_2	$b'_1/\bar{b}_1$	$b_1/\bar{b}_1=c_1/\bar{c}_1$	$c'_1/\bar{c}_1$
2	0.7500	0.7082	0.9443	1.1250	0.9288	0.7500
3	0.8888	0.8813	0.9916	1.0555	0.9703	0.8889
4	0.9375	0.9353	0.9976	1.0313	0.9838	0.9375
5	0.9600	0.9590	0.9990	1.0200	0.9900	0.9600

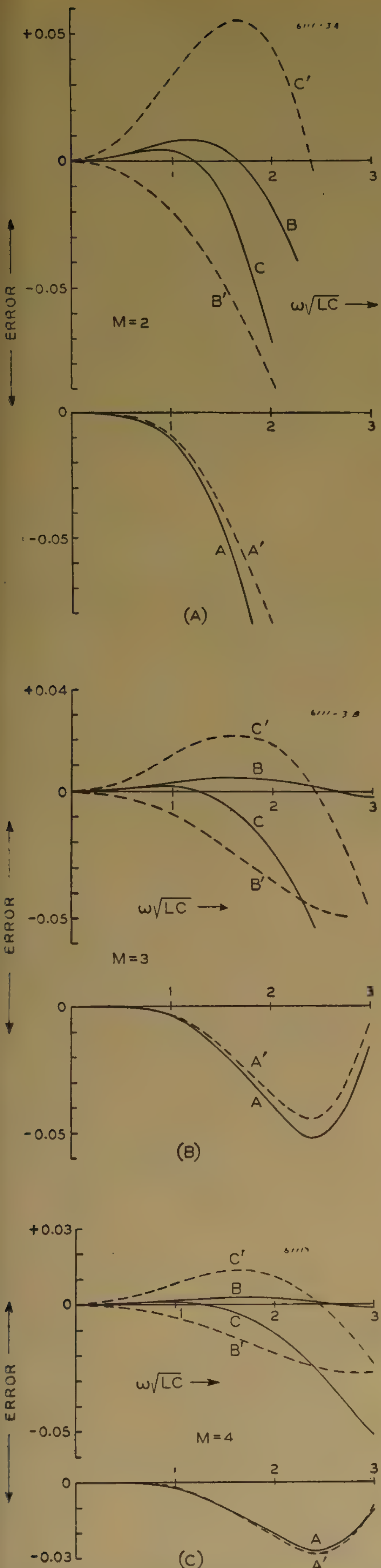

Figure 3. Performance of a dissipationless modified network

small significance. Therefore, although there will be differences, we should expect no appreciable difference between the new and old circuits when $|\alpha|$ is beyond the useful range. It is perhaps more correct to say that both circuits are equally bad in this extended range.

Other Modifications

Since the large error in a_2 may prevent the full realization of the gains made in b_1 and c_1, one may ask whether a more useful circuit might not be obtained if one could reduce the error of a_2 by allowing b_1 and c_1 to have errors of opposite algebraic sign. This question is treated in Appendix IV with results which are negative. It also is shown there that the minimum possible error in a_2 comes when the circuit is unmodified.

The above statements are with respect to variations of only the end elements. The possibility of improving a_2 by modifying the elements second from the end is considered in Appendix V. No improvement is found. The minimum possible error in a_2 still maintains when there is no modification at all. Furthermore, if b_1 and c_1 are again made equal, their minimum error comes when the shunt element has its uniform-network value and the terminal elements are modified in the manner already described (speaking of the series-element terminated case). Corresponding conclusions, of course, would follow if the shunt-element terminated case were treated. No analysis was made to determine the effects of modifying other elements of the network.

Conclusions

Within the limits afforded by the consideration of the end pair of elements, it has been shown that the best modification is that of the end elements alone. The specification for that modification is given below, in terms of the parameter $k=-2\rho$ which is introduced because ρ is negative. We have

$$\frac{Z'_T}{2M}+\rho Z_T=\frac{1-k(M-1)}{2M}Z_T \tag{21a}$$

$$\frac{Y'_T}{2M}+\rho Y_T=\frac{1-k(M-1)}{2M}Y_T \tag{21b}$$

for the series-element and shunt-element terminations, respectively. The other elements are chosen as if for a uniform network but with the series (or respectively shunt) elements each increased by the factor $(1+k)$.

It has been found possible to modify the terminations of a uniform ladder network to make it electrically more symmetrical in B and C (that is, the distinction between shunt-element and series-element termination is reduced) and electrically more nearly like its corresponding circuit of distributed characteristics, both effects being true only over a limited range of frequencies. Since the frequency range is independent of the number of sections and the improvements are small, the results are probably not of great practical importance. However, some improvement was found of a magnitude sufficient to make the new termination of value in an experimental application in which only a few sections are used. With many sections the need for accuracy would have to be extreme to make the new terminations useful.

In the discussion, emphasis has been placed on enhancing the similarity between the lumped network and a smooth line. Z_T and Y_T each can have any form, so any type of physical line is represented. Many forms of Z_T and Y_T, however, do not represent a realizable line at all. In such cases the electrical symmetry improvement, in the afore-mentioned sense, may be of some advantage.

In spite of the limitations on the practical value of the results, it is thought that they are worthwhile from the academic viewpoint alone. A theoretical survey is given, with definite proofs, of what can and cannot be done by varying both elements on each end of a uniform ladder network.

Appendix I

It is well known that only three of the four parameters of a 4-terminal network are independent and that when the network is symmetrical, only two are independent. The relation among them is well known[5] being, in our notation,

$$A^2-1=BC\alpha^2 \tag{22}$$

This is true for any symmetrical 4-terminal network, so the notation is perfectly general within that limitation. Expressing each side as a series in α^2 and carrying out the indicated multiplications give

$$a_0^2-1+2a_1a_0\alpha^2+(a_1^2+2a_0a_2)\alpha^4+\dots = b_0c_0\alpha^2+(b_1c_0+c_1b_0)\alpha^4\dots \tag{23}$$

Coefficients of like powers of α must be identical on the two sides so

$$a_0^2=1 \tag{24a}$$

$$2a_1a_0=b_0c_0 \tag{24b}$$

$$a_1^2+2a_0a_2=b_1c_0+c_1b_0 \tag{24c}$$

from which is obtained

$$a_0=1 \tag{25a}$$

$$a_1=\frac{b_0c_0}{2} \tag{25b}$$

These are generalizations of previous results making them applicable to any symmetrical network. It is seen that if $b_0=c_0=1$ and equations 25a–25b are substituted in equation 24c, the desired result, equation 11, is obtained. Equation 9 gives the necessary values for b_0 and c_0 for the case under consideration.

Appendix II

From a rearrangement of equation 14, one may write

$$6b_1=2\left(1-\frac{1}{2M^2}\right)\sqrt{1-\frac{M^2}{(2M^2-1)^2}}-1+\frac{1}{M^2} \quad (26)$$

The radical obeys the inequality

$$1-\frac{M^2}{2(2M^2-1)^2}-\frac{M^4}{2(2M^2-1)^4}<\sqrt{1-\frac{M^2}{(2M^2-1)^2}}<1-\frac{M^2}{2(2M^2-1)^2} \quad (27)$$

as will now be proved. The right-hand inequality is evident since the upper bound indicated is obtained if $M^4/4(2M^2-1)^4$ is added under the radical to form a perfect square. The writing of the proof for the left-hand half will be simplified if the letter W is used for the quantity $M/(2M^2-1)$. W^4 is greater than W^6 and W^8, because W is less than unity, so the following inequality may be formed:

$$\left(1-\frac{W^2}{2}-\frac{W^4}{2}\right)^2=1-W^2-\frac{3W^4}{4}+\frac{W^6}{2}+\frac{W^8}{4}$$
$$<1-W^2-\frac{3W^4}{4}+\frac{W^4}{2}+\frac{W^4}{4}$$
$$=1-W^2 \quad (28)$$

which is equivalent to the left-hand half of relation 27. The combination of formulas 26 and 27 yields a form which readily reduces to

$$1-\frac{1}{2(2M^2-1)}-\frac{M^2}{2(2M^2-1)^3}<6b_1<1-\frac{1}{2(2M^2-1)} \quad (29)$$

This is still too complex to be useful as an estimate, but it may be simplified by using the relations

$$2M^2-1=2\left(M^2-\frac{1}{2}\right)\geqq 2M^2\left(1-\frac{1}{8}\right)=\frac{7M^2}{4} \quad (30a)$$

and

$$(2M^2-1)^3=2^3\left(M^2-\frac{1}{2}\right)^3\geqq 2^3\left(M^2-\frac{1}{2}\right)^2$$
$$\geqq 2^3M^4\left(1-\frac{1}{8}\right)^2=\frac{49M^4}{8} \quad (30b)$$

which are true when $M\geqq 2$. Using these, and the relation $2M^2-1<2M^2$, appropriately in inequality 29 we get

$$1-\frac{4}{14M^2}-\frac{8M^2}{98M^4}<6b_1<1-\frac{1}{4M^2} \quad (31)$$

Relation 15 follows directly from this.

Appendix III

The results of Appendix II, the fact that $b_1=c_1$, and the simple form for c_1 given in equation 12b allow estimates of ρ and $1-2\rho$ to be obtained with ease. Without giving the details of the algebra, it is found that

$$\frac{M^2-18/49}{M^2-1}<1-2\rho<\frac{M^2-1/4}{M^2-1} \quad (32a)$$

and

$$\frac{31/98}{M^2-1}<-\rho<\frac{3/8}{M^2-1} \quad (32b)$$

For convenient usage in the subsequent work, these may be replaced by the inequalities

$$\frac{M^2-2/5}{M^2-1}<1-2\rho<\frac{M^2-1/4}{M^2-1} \quad (33a)$$

and

$$\frac{3/10}{M^2-1}<-\rho<\frac{3/8}{M^2-1} \quad (33b)$$

which are nearly as good.

Now consider the formula for a_λ. From equations 4a–4c it is obvious that $c'_{\lambda-1}=2\lambda a'_\lambda$, which may be used to simplify the expression for a_λ to

$$a_\lambda=a'_\lambda(1-2\rho)^{\lambda-1}[1+2\rho(\lambda-1)] \quad (34)$$

When $\lambda=2$, this reduces to $a_\lambda=a'_\lambda(1-4\rho^2)$, which reduces to relation 18 when bounds of ρ are taken from inequality 33b. A general estimate may be obtained for $\lambda\geqq 3$ by using both parts of relation 33 and whichever value of λ (3 or M) enhances the limit in question. This procedure yields inequality 19a.

Proceeding to the derivation of inequality 19b, we find complication because of the complexity of the expression for b_λ. From equations 4a–4c, it follows that $a'_\lambda=b'_\lambda\times(2\lambda+1)(2M^2)/(2M^2+\lambda)$, and $c'_{\lambda-1}=b'_\lambda\times(2\lambda)(2\lambda+1)(2M^2)/(2M^2+\lambda)$. These, when substituted in equation 8b, yield

$$b_\lambda=b'_\lambda(1-2\rho)^{\lambda-1}\left\{(1-2\rho)^2+[\rho^2(2\lambda+1)\times(2\lambda)+2\rho(1-2\rho)(2\lambda+1)]\frac{2M^2}{2M^2+\lambda}\right\} \quad (35)$$

An upper bound of b_λ will first be found. The first term in the brackets is positive and the second is negative. This necessitates the use of upper bounds for all components of the former and lower bounds for all components of the latter. Since $2M^2/(2M^2+\lambda)$ is a factor of each, it is subject to this general statement in its two roles as a multiplier. Satisfactory upper and lower bounds for $-\rho$, ρ^2, and $1-2\rho$ are given in relations 33. When $2M^2/(2M^2+\lambda)$ multiplies the first term, it may be replaced by 1, while in the other case it may be replaced by the lower bound $M/(M+1)$. Also $(2\lambda+1)(2\lambda)$ is less than $(2M+1)(2M)$, and $(2\lambda+1)\geqq 5$ if $\lambda\geqq 2$. Omitting the details of the algebra, it is found that

$$b_\lambda<b'_\lambda\frac{(M^2-1/4)^{\lambda-1}}{(M^2-1)^{\lambda+1}}\left(M^4-2M^3-\frac{15M^2}{16}+\frac{105M}{32}+\frac{1}{16}\right) \quad (36)$$

When $M\geqq 4$ the quantity represented by the three terms on the right of the above parenthesis is less than one, so if this restriction is imposed on M, the entire parentheses on the right may be replaced by $(M^2-1)^2$, which conveniently cancels against part of the denominator, leaving

$$b_\lambda<b'_\lambda\left(\frac{M^2-1/4}{M^2-1}\right)^{\lambda-1} \quad (37)$$

The upper bound is treated in a similar fashion. When $\lambda\geqq 2$, the quantity $(2\lambda+1)\times(2\lambda)$ may be replaced by the lower bound 20; in the other term $(2\lambda+1)$ may be replaced by the upper bound $(2M+1)$; and $2M^2\div(2M^2+\lambda)$ may be replaced by the upper and lower bounds 1 and 1/2 in appropriate locations. Proceeding as before, one obtains, after simplifications,

$$b_\lambda>b'_\lambda\frac{(M-2/5)^{\lambda-1}}{(M^2-1)^{\lambda+1}}\left(M^4-\frac{3M^3}{2}-\frac{31M^2}{20}+\frac{3M}{8}+\frac{499}{400}\right) \quad (38)$$

When $M\geqq 4$, the expression in the parentheses on the right is greater than $M^4-2M^3+2M-1=(M-1)^2(M^2-1)$, which if substituted cancels partially against the denominator to yield

$$b_\lambda>b'_\lambda\left(\frac{M-2/5}{M^2-1}\right)^{\lambda-1}\left(\frac{M-1}{M+1}\right) \quad (39)$$

Inequalities 37 and 39 combine to give inequality 19b.

Recalling the formula for c_λ, we see that it already is taken care of by the estimate of $1-2\rho$. Relation 19c follows directly.

Appendix IV

It is evident from the analysis leading up to equation 13 that b_1 and c_1 each cannot be equal to 1/6. However, one might suppose that b_1 and c_1 could be given errors of opposite sign to make $b_1+c_1=1/3$. Writing this out from equations 8b–8c, we obtain

$$\rho(1-2\rho)+\frac{1}{6}\left(1+\frac{1}{2M^2}\right)(1-2\rho)^2+\rho^2+\frac{1}{6}(1-2\rho)\left(1-\frac{1}{M^2}\right)=\frac{1}{3} \quad (40)$$

the solution of which is

$$\rho^2=-\frac{1}{4(M^2-1)} \quad (41)$$

The indicated imaginary value for ρ is physically impossible using passive elements because it must hold over a range of frequencies.

Let us now find the condition on ρ that will give a minimum error in a_2. At such a value the derivative of b_1+c_1 with respect to ρ will be zero. Differentiating the left side of equation 40 and setting the result equal to zero give, after reduction,

$$\frac{2\rho}{3}\left(\frac{1}{M^2}-1\right)=0 \quad (42)$$

Since $M=1$ never occurs, the solution is $\rho=0$. It is evident, since there is only one root and since the error can never be zero, that

this corresponds to a minimum and not a maximum error.

Appendix V

The analysis for the changing of two elements on each end of the network will proceed with reference to the circuit of Figure 4. This is built up by starting with the mid-shunt terminated network between lines P-P, with a subsequent double application of the modification principle already described in the text. The original network has a total impedance and admittance, respectively, of Z'_T and Y'_T. We first add a shunt element of magnitude βY_T^*, where Y_T^* will indicate the new total admittance. The transformation, in similarity with equations 5a–5d, involves the quantities

$$\alpha^*=\sqrt{Z_T^*Y_T^*} \tag{43a}$$

$$Z'_T=Z_T^* \tag{43b}$$

$$Y'_T=(1-2\beta)Y_T^* \tag{43c}$$

$$\alpha'=\alpha^*\sqrt{1-2\beta} \tag{43d}$$

As has been pointed out in the text, equations 8a–8c will give the coefficients for the new circuit if the equations for b_λ and c_λ are interchanged, due attention being given to differences in notation. Specifically, a_λ, b_λ, and c_λ become, respectively, a_λ^*, c_λ^*, and b_λ^*, while a'_λ, b'_λ, and c'_λ still are given by equations 4a–4c. The circuit between the lines Q-Q is then in a condition to receive a series element on each end, giving it the appearance of a modified mid-series terminated network. Let the added element on each end be σZ_T, and let the following additional notational changes take place:

$$\alpha=\sqrt{Z_TY_T} \tag{44a}$$

$$Z_T^*=(1-2\sigma)Z_T \tag{44b}$$

$$Y_T^*=Y_T \tag{44c}$$

$$\alpha^*=\alpha\sqrt{1-2\sigma} \tag{44d}$$

Equations 8a–8c again may be applied, this time as they stand, with a'_λ, b'_λ, and c'_λ replaced by the corresponding starred coefficients. Thus, unprimed coefficients apply to the over-all circuit of Figure 4. For simplicity of notation, M-1 will be replaced by N until the results are achieved. It will be necessary to consider only the four coefficients b_0, b_1, c_0, and c_1, which, as derived by the foregoing plan of action, are

$$b_0=2\sigma a_0^*+b_0^*(1-2\sigma)=1 \tag{45a}$$

$$\begin{aligned}b_1&=2\sigma(1-2\sigma)a_1^*+(1-2\sigma)^2b_1^*+\sigma^2c_0^*\\&=\sigma(1-2\sigma)+(1-2\sigma)^2(1-2\beta)c'_1+\sigma^2\end{aligned} \tag{45b}$$

$$c_0=c_0^*=1 \tag{45c}$$

$$\begin{aligned}c_1&=(1-2\sigma)c_1^*\\&=(1-2\sigma)[2\beta a'_1(1-2\beta)+b'_1(1-2\beta)^2+c'_0\beta^2]\end{aligned} \tag{45d}$$

The simplifications which already have been included in the writing of these equations come from the facts that $a_0^*=b_0^*=c_0^*=1$, and $a_1^*=1/2$, in similarity with equations 9 and 10. Since the correctness of b_0 and c_0 assures the correctness of a_1, as before, our attention may be confined to b_1 and c_1. The algebraic work may be expedited by changing the notation to

$$x=1-2\beta \tag{46a}$$

$$y=1-2\sigma \tag{46b}$$

The reduced formulas for b_1 and c_1 are obtained by taking a'_1, b'_1, c'_0, and c'_1 from equations 4a–4c, with M replaced by N. This gives

$$b_1=\frac{1}{4}-\frac{y^2}{4}+\frac{1}{6}\left(1-\frac{1}{N^2}\right)xy^2 \tag{47a}$$

$$c_1=\left[\frac{1}{4}-\frac{1}{12}\left(1-\frac{1}{N^2}\right)x^2\right]y \tag{47b}$$

We first will see if b_1 and c_1 each can be made equal to 1/6, or at least if their sum can be 1/3 so that a_2 will be correct. Accordingly, assume $b_1=1/6+\epsilon$ and $c_1=1/6-\epsilon$, which represents both conditions, since ϵ

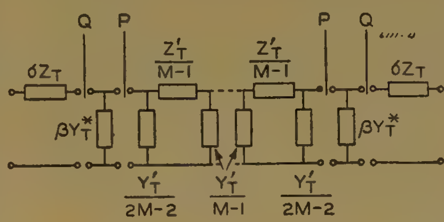

Figure 4. Two-element modification

may be zero. The result of these substitutions and the subsequent combination of equations 47a–47b is a quadratic in ϵ, as follows:

$$144G\epsilon^2+(12F-48G)\epsilon+(4G-F)=0 \tag{48}$$

where

$$G=G(N,x)=3-2\left(1-\frac{1}{N^2}\right)x \tag{49a}$$

$$F=F(N,x)=9-6\left(1-\frac{1}{N^2}\right)x^2+\left(1-\frac{1}{N^2}\right)^2x^4 \tag{49b}$$

The discriminant of equation 48 is easily computed to be $144F(F-4G)$. It must be positive if a real solution is possible, so we are interested in

$$F-4G=\left(\frac{N^2-1}{N^2}\right)\left(\frac{N^2-1}{N^2}x^4-6x^2+8x-\frac{3N^2}{N^2-1}\right)=\frac{N^2-1}{N^2}H(N,x) \tag{50}$$

Before investigating the positiveness of $H(N, x)$, it should be noted that to have a physically realizable circuit it is necessary that $-1/2(N-1)<\beta<1/2$. Outside of this range the end shunt elements either would vanish completely, thus changing the number of network sections, or would be greater than half the total admittance. The corresponding range of x is $0<x<N/(N-1)$. If $H(N, x)$ is computed for $N=\infty$ throughout the largest possible range of x, it will be found to be everywhere nonpositive. Furthermore, as may be easily shown, $\partial H\times(N,x)/\partial N$ is always positive, so $H(N,x)$ is less than $H(\infty, x)$ and is therefore always negative in the pertinent range. Thus there is no real solution of equation 48, so b_1 and c_1 cannot be real and also have 1/3 as their sum.

Following a procedure similar to that used in the simpler case, we next investigate whether the error in a_2 can be made a minimum. The total differential of b_1+c_1 is

$$d(b_1+c_1)=\left(\frac{\partial b_1}{\partial x}+\frac{\partial c_1}{\partial x}\right)dx+\left(\frac{\partial b_1}{\partial y}+\frac{\partial c_1}{\partial y}\right)dy \tag{51}$$

The smallest possible error will occur when dx and dy are independent and thus when their coefficients are each zero. As may be seen easily by carrying out the necessary differentiation, equating the coefficient of dx to zero gives

$$\frac{1}{6}\left(1-\frac{1}{N^2}\right)(y^2-xy)=0 \tag{52}$$

from which one obtains the condition $y=x$, since $N\neq 1$. With the substitution of x for y in the derivatives of the coefficient of dy, the setting of this equal to zero gives

$$x^2-\frac{2N^2}{N^2-1}+\frac{N^2}{N^2-1}=\left(x-\frac{N}{N-1}\right)\times\left(x-\frac{N}{N+1}\right)=0 \tag{53}$$

The $N/(N-1)$ solution gives a value of β which wipes out the shunt element on each end and is found to correspond to a maximum error. The other solution gives $\beta=1/2(N+1)$. The corresponding complete end element is $Y_T[(1-2\beta)/2N+\beta]=Y_T/(N+1)=Y_T/M$. This means that all shunt elements will be the same. Also, the series element is $\sigma Z_T=\beta Z_T$, since $y=x$, which gives the value $Z_T/2M$, which is a mid-series termination. This should have been expected, since the case for uniform shunt elements has already been treated.

Now suppose we make $b_1=c_1$. This makes y and x interdependent, so dy and dx in equation 51 are no longer independent. Also, since b_1 and c_1 are to be equal as x and y vary, it is only necessary to take the derivative of one of them. dy/dx may be obtained from the implicit function $b_1-c_1=0$ as

$$\frac{dy}{dx}=-\frac{\partial(b_1-c_1)/\partial x}{\partial(b_1-c_1)/\partial y} \tag{54}$$

Thus

$$\frac{db_1}{dx}=\frac{\partial b_1}{\partial x}+\frac{\partial b_1}{\partial y}\frac{dy}{dx}=\frac{\frac{\partial b_1}{\partial x}\left(\frac{\partial b_1}{\partial y}-\frac{\partial c_1}{\partial y}\right)-\frac{\partial b_1}{\partial y}\left(\frac{\partial b_1}{\partial x}-\frac{\partial c_1}{\partial x}\right)}{\frac{\partial b_1}{\partial y}-\frac{\partial c_1}{\partial y}} \tag{55}$$

The denominator of equation 55 cannot be infinite, so we set the numerator equal to zero, giving, after reduction,

$$-\frac{\partial b_1}{\partial x}\frac{\partial c_1}{\partial y}+\frac{\partial b_1}{\partial y}\frac{\partial c_1}{\partial x}=0 \tag{56}$$

The substitution of the expressions for these derivatives gives an equation from which y can be eliminated. The remaining equation in x is identical with equation 53. Thus $x=N/(N+1)$ is again the solution, but of course y is no longer equal to x. It is not necessary to compute y because it has been

Geometric Mean Distances for Rectangular Conductors

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THE USE of geometric mean distance for calculating reactance of parallel conductors is increasing, so much so that the self geometric mean distance often is tabulated in catalogs of heavy conductors along with other characteristics, such as resistance, current capacity, and weight.

The logarithm of the geometric mean distance between two areas or cross sections is defined as the average of the logarithms of all possible distances from points on one area to those on the other, and the logarithm of the self geometric mean distance of an area is the average of the logarithms of all possible distances between two points on the area.[1]

The inductance, as computed by the geometric mean distance method, is the inductance at zero frequency. Metal of uniform conductivity, such as copper, is taken to have constant current density, or more generally, the current in a conductor is taken to be proportional to the conductance of that conductor, as it would be with direct current.

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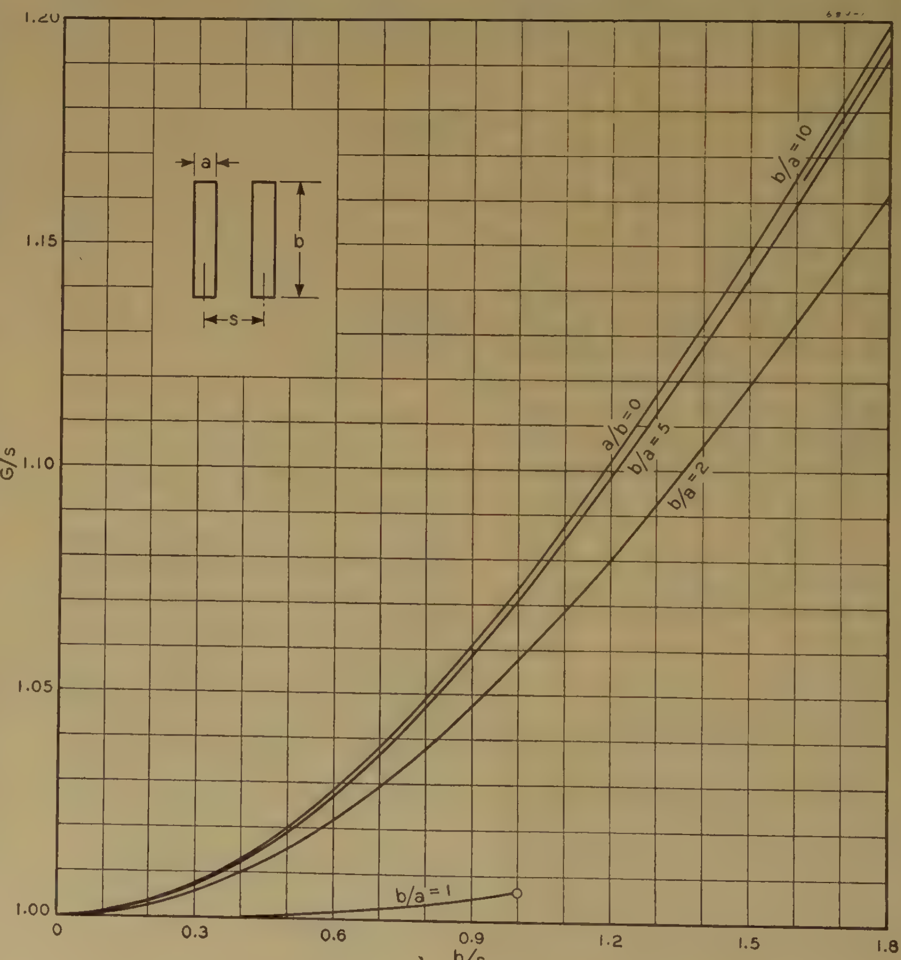
The self geometric mean distance of an area is sometimes called the geometric mean radius of the area, although it is not a radius nor an average of radii. It is, as the definition states, the mean distance (geometric, as distinguished from arithmetic) of all possible distances between two points on the area, and of course, most of these are not radii. It seems better to use a name which is in agreement with the facts of the case, rather than one which is practically no more convenient and which does not describe the quantity truly.

Formulas for the geometric mean distance of geometrical shapes can be obtained by a process of integration.[1] However, the formula for two rectangles, even when the problem is simplified by specifying two duplicate rectangles placed opposite each other, fills half a page. For engineering purposes, curves of reactance values for rectangular conductors can be used.[2] See also Tables I–III in a paper published in 1929 by Professor F. W. Grover.[3]

In this paper are given curves for geometric mean distance between rectangular areas which will give a somewhat more direct solution for general types of problems, such as those involving polyphase circuits or conductors in parallel.

In circuits or portions of circuits where the parallel conductors are not transposed, the reactive voltage drop in the various conductors is different. It may be computed by counting flux caused by each conductor, up to a certain large distance u. In a complete system of parallel conductors in the steady state, the sum of all the currents is zero, that is there is as much return current as going current in the group. The quantity u cancels out as it is associated with all of the conductors in turn (see equation 2). Therefore, the result is the same,

Figure 1. Geometric mean distance of rectangular areas, face-to-face position



shown that when x has this value the shunt elements are all uniform. This means that the determining of y is equivalent to the finding of ρ in the problem as originally treated.

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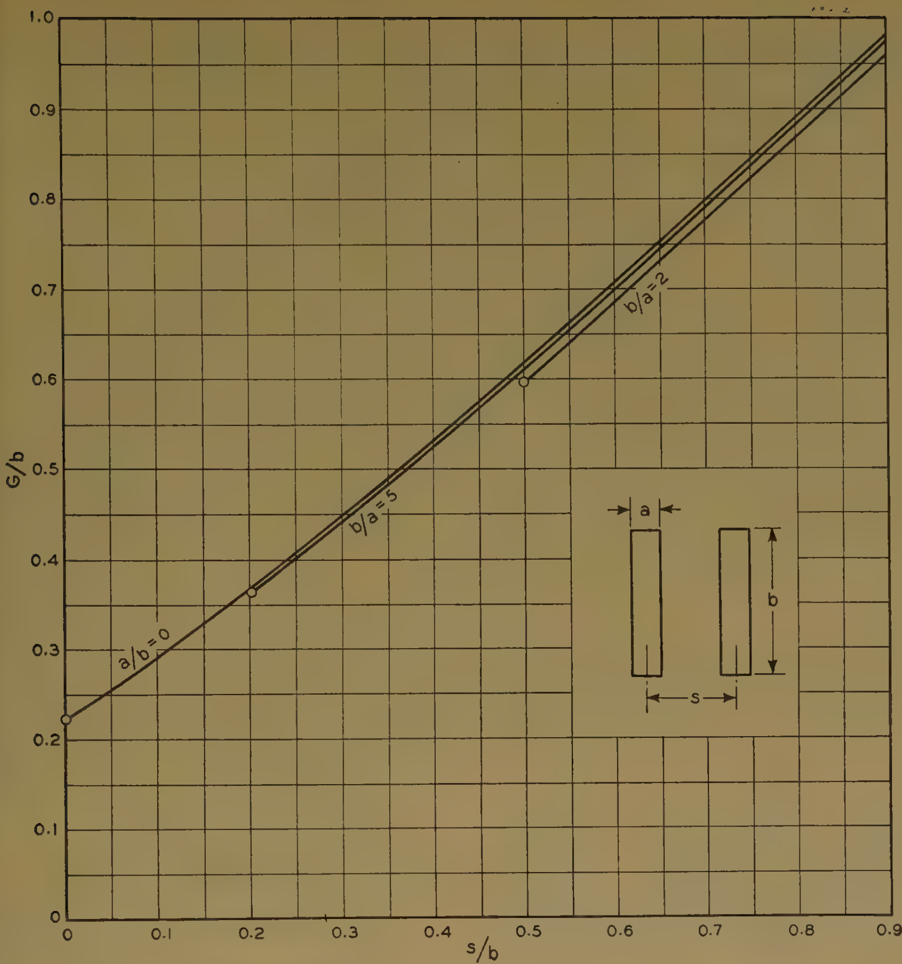


Figure 2. Geometric mean distance of rectangular areas, face-to-face position, close spacing

no matter how large u may be. The currents, as well as the conductor arrangement, may be unbalanced or balanced and are in amperes.

If R_A is the resistance per centimeter of conductor A of a group of long parallel conductors A, B, C, . . ., if G_{sA} is its self geometric mean distance, if G_{AB} is the geometric mean distance of the section of A from that of B, and so forth, the voltage drop in conductor A is

$$I_A R_A + j4\pi f \times 10^{-9}\left[I_A \log_\epsilon \frac{u}{G_{sA}} + I_2 \log_\epsilon \frac{u}{G_{AB}} + I_3 \log_\epsilon \frac{u}{G_{AC}} + \ldots \right]$$

volts per centimeter (1)

where f is the frequency. Putting

$$(I_A + I_B + I_C + \ldots) \log_\epsilon u = 0 \qquad (2)$$

u being a finite constant, although as large as desired, we have the reactive voltage drop in conductor A:

$$= -j4\pi f \times 2.303 \times 10^{-9}[I_A \log_{10} G_{sA} + I_B \log_{10} G_{AB} + I_C \log_{10} G_{AC} + \ldots]$$

volts per centimeter (3)

or

$$= -j0.0529\,[I_A \log_{10} G_{sA} + I_B \log_{10} G_{AB} + I_C \log_{10} G_{AC} + \ldots]$$

volts per 1,000 feet at 60 cycles (4)

Note that A, B, and C necessarily do not mean different phases, but merely different conductors.

Equations 3 and 4 are general formulas for the reactive voltage drop in any one of a group of long parallel conductors whose currents are given. The conductors may be of different phases or they may be branches in parallel.

The self geometric mean distance of a rectangular sectional area $a \times b$ is given within 0.13 per cent by the formula

$$G_s = 0.2234(a+b) \qquad (5)$$

Actually, the numerical term varies from 0.2231 to 0.2237 for different ratios of a to b.[4]

The geometric mean distance between two equal rectangular areas, placed opposite each other with the sides parallel, may be read from the curves shown in Figures 1, 2, or 3, depending on the ratios of the dimensions. For the use of these curves see the examples.

The geometric mean distance between two equal rectangular areas, placed obliquely and not very close together, may be taken as the distance between their centers for a first approximation. The percentage error resulting from this procedure may be estimated by reference to Figures 1 and 3. For instance, if the distance between centers of two comparatively thin straps is 1.25 times the strap width, then from Figure 1, for the parallel plane position, $b/s = 0.8$ and G is five per cent greater than s. From Figure 3, for the edgewise position, $a/s = 0.8$ and G is six per cent less than s. Then for an oblique position, with the afore-mentioned distance between centers, the error in taking $G = s$ may be expected to be numerically less than plus or minus six per cent. However, this result does not mean six per cent error in reactance, since log G is involved.

In either equation 3 or 4, the distances G_{sA}, G_{AB}, G_{AC} . . . may be all in centimeters or all in inches (or in feet), so long as they are all in the same unit.

When there is complete transposition so that each conductor occupies all the positions in succession to an equal extent, the geometric mean distance method may be extended to give a convenient calculation for the average voltage drop. For this calculation, values of geometric mean distance of rectangles taken from Figures 1, 2, and 3 are useful, as well as in problems where equations 3 and 4 are used.

Precise values of $\log_\epsilon$ geometric mean distance are given for equal squares in many relative positions.[5] By combining these squares into rectangles of various shapes and positions, the precise value of the logarithm of the geometric mean distance was computed for a considerable number of cases of rectangles opposite each other. From these determinations and from values of G_s (equation 5), reactance values in micro-ohms per foot at 60 cycles were obtained. Formulas for infinitely thin rectangular conductors, that is for $b/a = 0$ and $a/b = 0$, are given[4] and curves of reactance values were plotted.[2]

In the present paper, instead of values of loop reactance, values of the geometric mean distances themselves are plotted. Their use has been described and is illustrated in the examples. For these curves a number of additional values have been determined.

Example 1

Find the reactive voltage drop in a single-phase ventilated bus bar circuit made up of four straps of size 1/8 by 3 inches, side by side in vertical planes. Each bus bar consists of two straps at $^1/_2$-inch centers. The distance between centers of bus bars is $2^1/_2$ inches. The cur-

rent of the circuit is 900 amperes, 60 cycles.

Let the four straps be called A, B, C, and D. Then the center-to-center distance from A to B is $1/2$ inch, 2 inches from B to C, and $1/2$ inch from C to D. A and B are connected in parallel. Neglecting circulating currents, A and B will each carry $+450$ amperes, and C and D each -450 amperes.

For computing the voltage drop in A by equation 4, using logarithms to base 10,

$G_s = 0.2234 \times 3.125 = 0.698$ inch
$\log 0.698 = \bar{1}.8439 = -0.1561$

For G_{AB},

$s/b = 0.5/3 = 0.167,\ b/a = 24$

From Figure 2,

$G_{AB} = 0.343 \times 3.0 = 1.029$
$\log 1.029 = 0.01242$

For G_{AC},

$b/s = 3/2.5 = 1.20$

From Figure 1,

$G_{AC} = 1.101 \times 2.5 = 2.752$
$\log 2.752 = 0.4396$

Alternatively, by Figure 2,

$G_{AC} = 0.917 \times 3 = 2.751$

Similarly,

$$
\begin{aligned}
&\log G_{AD} = 0.508\\
&I_A \log G_s = 450 \times (-0.1561) = -\ 70\\
&I_B \log G_{AB} = 450 \times 0.01242 = \qquad +6\\
&I_C \log G_{AC} = -450 \times 0.439 = -198\\
&I_D \log G_{AD} = -450 \times 0.508 = \underline{-228}\\
&\qquad\qquad\qquad\qquad\qquad\quad -496+6\\
&-496+6 = -490
\end{aligned}
$$

By equation 4, the reactive drop in A for 70 feet is

$j\,0.0529 \times 490 \times 0.070 = j\,1.81$ volts

In a similar way, the drop in B is found to be

$$-j\,0.0529 \times 0.070(-70+6-163-198) = j\,1.57 \text{ volts}$$

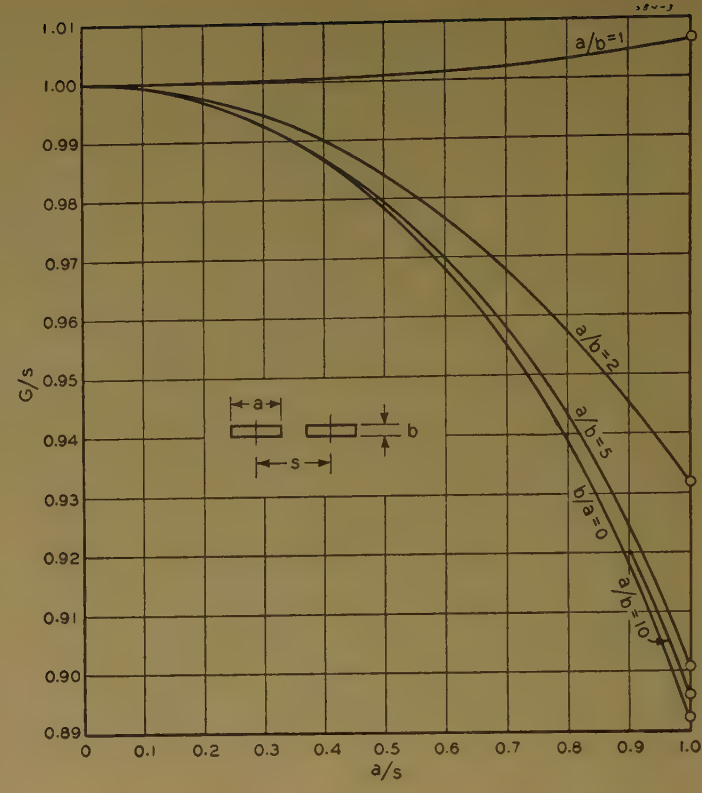
The difference in these computed voltage drops in A and B may be said to result in an unbalance of current, which may be computed if desired.

The result of this problem agrees with the results of previously computed problems.[2]

Example 2

Find the reactive drop in each conductor of a 3-phase circuit consisting of three rectangular conductors (one conductor per phase) in parallel planes, the distance between centers being $1\,1/2$

Figure 3. Geometric mean distance of rectangular areas, edgewise position



inches. The conductors are $1/4 \times 3$ inches and each is 100 feet long. The current is 750 amperes, 60 cycles.

Let the three conductors be called A, B, and C, and let

$I_A = 750$ amperes
$I_B = 750(-0.5 + j\,0.866)$ amperes
$I_C = 750(-0.5 - j\,0.866)$ amperes

By equation 5,

$G_s = 0.2234\ (0.25+3) = 0.726$ inch
$\log 0.726 = \bar{1}.8609 = -0.1391$, omitting the subscript 10 in $\log_{10}$

For G_{AB},

$s/b = 1.5/3 = 0.5,\ b/a = 12$

From Figure 2,

$G_{AB} = 0.616 \times 3.0 = 1.848$ inches
$\log 1.848 = 0.2667$

For G_{AC},

$b/s = 3/3 = 1,\ b/a = 12$

From Figure 1,

$$
\begin{aligned}
&G_{AC} = 1.072 \times 3.0 = 3.216 \text{ inches}\\
&\log 3.216 = 0.5073\\
&I_A \log G_s = 750(-0.1391) &&= -104\\
&I_B \log G_{AB} = (-375 + j650) \times 0.2667 &&= -100 + j173\\
&I_C \log G_{AC} = (-375 - j650) \times 0.5073 &&= \underline{-190 - j330}\\
& &&\quad -394 - j157
\end{aligned}
$$

By equation 4, the reactive drop in A for 100 feet is

$$-j\,0.0529 \times 0.10(-394 - j157) = -0.83 + j2.09 \text{ volts} = 2.24 \text{ volts}$$

For the reactive drop in conductor B,

$$
\begin{aligned}
&I_B \log G_s = (-375 + j650) \times (-0.1391) &&= 52 - j\,90\\
&I_C \log G_{BC} = (-375 - j650) \times 0.2667 &&= -100 - j173\\
&I_A \log G_{AB} = 750 \times 0.2667 &&= \underline{200}\\
& &&\quad 152 - j263
\end{aligned}
$$

By equation 4, the reactive drop in B is

$$-j\,0.0529 \times 0.10(152 - j263) = -1.40 - j\,0.81 = 1.62 \text{ volts}$$

The reactive drop in conductor C has the same numerical value, 2.24 volts, as in A.

The result of this problem agrees with already proved similar problems.[2]

It is a little more straightforward to use geometric mean distances and equation 4 as in this paper, than to rearrange the problem so as to use reactance values of single loops. The geometric mean distance method is applicable also to a wide range of reactance problems.

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The Frequency Response of Automatic Control Systems

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Synopsis: This paper illustrates the advantages of the frequency response approach to the analysis of automatic control systems, as contrasted to the analytic solution of transient responses. At the same time the equivalence of information available from the two approaches is demonstrated. A numerical example is given based upon a torque amplifier using a motor-generator type control.

THE purpose of this paper is to show how the concept of frequency response is being applied analytically and experimentally in the design of automatic control systems such as servomechanisms. This concept long has been used in the radio and telephone arts to express the fidelity of response of equipment, and as a result a considerable background of established mathematics and experimental procedure has been made available to instrument and automatic control engineers.

A transient analysis of the response of an automatic control to some standard type of disturbance is an accepted procedure, but with the more complex systems it becomes unwieldy, and useful design criteria become increasingly difficult to develop. A frequency analysis based on the mathematics of the Fourier series and the Fourier integral provides information equivalent to the transient analysis but with much less calculation and with more easily interpretable results. This paper will show this equivalence and the mathematical procedures involved in the design of a servomechanism.

Typical Servomechanisms

A servomechanism is a system in which some variable quantity is controlled to follow a definite function of some other varying quantity. Typical servomechanisms are

1. A self-balancing potentiometer[1] in which the controlled variable is the mechanical position of the pen and slide wire, and the desired position is proportional to the varying quantity, namely the voltage under measurement.

2. A machine tool pattern follower for which the controlled variable is the tool position, and the desired position is proportional to the varying shape of the pattern as the tool moves along.

3. A data transmission system[2] which reproduces a mechanical motion at a remote point, often with greatly increased force.

4. A highly degenerative feed-back amplifier[3] in which the output voltage faithfully reproduces a varying input voltage. The recognition of the similarity between servomechanisms and feed-back amplifiers makes available to the automatic control engineer many valuable analytical tools developed by communication engineers.

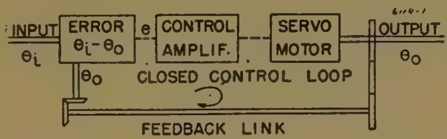

Figure 1. Torque amplifier

Figure 1 is a block schematic of the components of a torque amplifier which reproduces the motion of the input shaft and drives a heavy load, without placing any appreciable load on the input shaft. The load is driven by a motor controlled by a signal or motion from a device which measures the difference or error between the input and output shafts. This error-measuring device is the only load on the input shaft and may be made quite light. All of the examples listed are similar in that they all comprise an input, an output, an error-measuring device, an amplifier, and a power unit which drives the load to bring the error to zero.

The Transient Analysis

All systems which take the form of a closed loop as in Figure 1 have a tendency to be unstable and hunt, or at least to perform damped oscillations about their zero error position. The first problem in proper servomechanism design is to minimize these oscillations after any transient disturbance. The second problem of proper design is to minimize errors which occur under steady conditions, such as a constant velocity or acceleration motion of the input shaft, and errors caused by the loads on the output shaft. Usually the solution must be a compromise between the best solutions for these two problems.

One procedure for studying the oscillatory character of a system analytically is to assume a motion for the input and then calculate the resultant transient response of the output. A typical input is a suddenly applied constant velocity as is shown in Figure 2 along with a typical resulting output motion.

With the assumption that the system is linear over the region under investigation, any one of several classical, operational, or transform methods[4] for solving linear differential equations with constant coefficients may be used for calculating such a transient response. Whatever method is used, there are four essential steps in the solution.

1. The differential equations for the system are set up.

2. The characteristic equation of the above set is obtained. This equation has the form

$$a_n p^n + a_{n-1} p^{n-1} + \ldots a_1 p + a_0 = 0$$

in which the coefficients contain the various parameters of the system, such as amplifier gain, inertia, friction, and circuit constants.

3. The characteristic equation is solved for its roots. The roots will be real or in conjugate complex pairs. $p_n = \sigma_n \pm j\omega_n$

4. The final solution contains a term of the form $Ae^{\sigma t}\cos(\omega t+\alpha)$ for each complex root, an exponential term for each real root, and a particular solution which depends on the input motion. The coefficients A and phase angles α will depend upon the initial conditions of the system.

In this solution the rate of decay of the oscillatory terms is determined by the real parts of the complex roots, namely the σ_n's. If these are large and negative, the oscillations will damp quickly. If one is small, then the damping will be slow. If any one is positive, the system will be unstable and oscillations will increase.

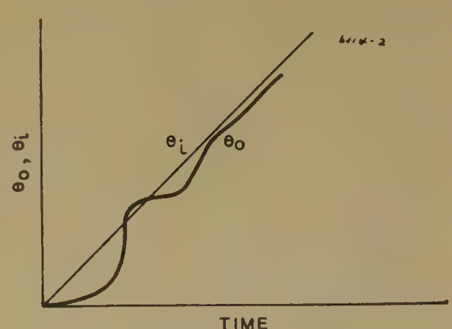

Figure 2. Transient response to an input velocity

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The solution of the characteristic equation is the key point in all the methods of transient analysis. If the degree n is five or more, only a numerical solution may be obtained and then only by such laborious calculations that an engineer often cannot afford the necessary time. This situation is accentuated by the fact that even after a solution is found there is often no feasible way to tell what parameters must be changed to improve the system because the system parameters are so completely intermixed in the coefficients a_n. A later illustration will bring out this point.

Stability Criteria

Instead of solving the characteristic equation, a compromise often is resorted to which merely tells whether the system is stable or unstable. Hurwitz or Routh criteria[4] when applied to the coefficients a_n will tell whether there are any roots with positive real parts. This may be an unsatisfactory compromise, since many stable systems damp very slowly and the stability criteria do not show what changes are needed for improvement.

The System Transfer Function

The frequency response method of analysis eliminates the necessity for solving the characteristic equation and brings out with great clarity the effect of each parameter on the performance of the system. For this method to be valuable, it is necessary to show first that the obtainable results give information which is at least roughly equivalent to that obtained from a transient solution.

Consider the input and output motions as shown in Figure 2. The transient completely dies out after a certain period of time, and if a new similar disturbance is introduced after this time, the response will repeat. This may be repeated again and again so that the input and the response will appear as in Figure 3. As is well known, both of these periodic time functions may be analyzed into a Fourier series of sinusoidal terms of frequencies which are multiples of $1/T$. By including a sufficient number of terms in the series, the motions may be approximated arbitrarily closely.

Corresponding to each sinusoidal term which is a component of the input motion there is a term of the same frequency which is a component of the output motion. The two corresponding terms at each frequency will differ in amplitude and in phase angle between the input and the output. The ratio of the output amplitude to the input amplitude is known as the magnitude of the system transfer function, and since it is a function of frequency, it may be plotted as in Figure 4. Similarly, the difference of the input and output phase angles is called the phase of the transfer function and may be plotted also as in Figure 4.

Since all the input frequency components add together to form the input transient and all the output components add together to form the output transient, the transfer function which tells how these components are modified must contain all the information available from the system differential equation. The proper interpretation of curves such as those in Figure 4 will yield this information. This interpretation is discussed immediately following, while the calculating procedure is deferred until later.

Interpretation of the Transfer Function

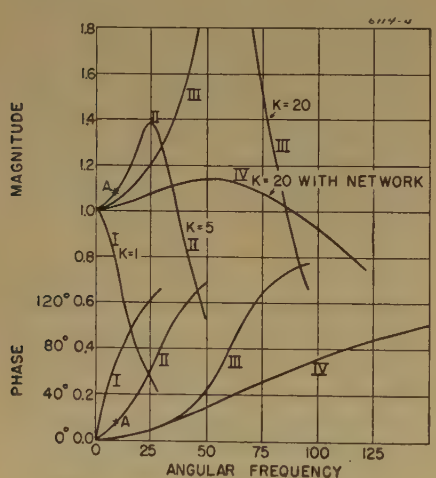
The ideal transfer function from the standpoint of perfect following of the input would be one which does not in any way change the components. This would occur with a magnitude of unity and a phase difference of zero over the whole frequency range. Actual transfer functions differ from the ideal in three important aspects.

1. Certain frequencies are accentuated or amplified as illustrated by the frequencies near $\omega=25$ radians per second for curve II of Figure 4. This peaking of these frequencies is called resonance. A transient disturbance will result in a damped oscillation at a frequency near the resonant frequency as shown in a corresponding transient response curve, II of Figure 5. There is a correlation between the height of the resonance peak and the time required for the transient oscillation to die out, but it is not generally expressible in quantitative form.

The allowable height of a resonance peak can be specified only from experience with systems of similar application, but this is also true of any specification of desired damping for a transient response. For data transmission systems such as the one described below, a peak of about 1.2 is a good maximum value for high performance systems, but peaks as high as 1.5 are sometimes acceptable.

2. The higher frequencies always are attenuated; that is, the magnitude of the transfer function approaches zero as the frequency is increased. This results in the output being more or less sluggish and less harp than the input motion. This is illustrated by curves I and IV of Figures 4 and 5. The system which passes more of the high frequencies has the faster motion. Two systems which have exactly the same form of transfer function but with different frequency scales will have exactly the same form of transient response but with a time scale which is inverse to the frequency scale. Thus, by doubling the frequency response of a system the transient response time is halved.

Another aspect of the frequency bandwidth concept is that the usual input motions predominate in low frequencies. For example, even in the unattainable periodic square wave of position in which the input periodically jumps forward a certain amount and then back, the amplitude of the frequency components drops off as $1/\omega$. For a periodically positive and negative velocity which is less sharp, the amplitudes drop off



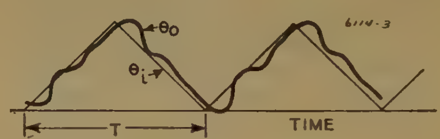
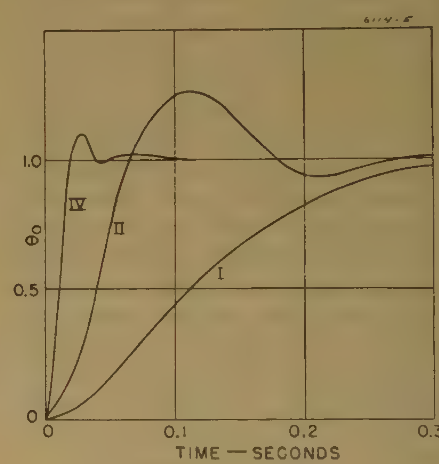


Figure 3 (above). Response to a periodic input

Figure 4 (left). System transfer functions

Figure 5 (right). Response to a unit step input



as $1/\omega^2$, and for a periodically positive and negative acceleration, the amplitudes are proportional to $1/\omega^3$.

Since most input motions predominate in the low frequencies, more attention should be paid to the behavior of the transfer function at low frequencies. As an example, a resonance peak at a high frequency acceptably may be somewhat greater than a peak at a low frequency, since it is less likely to be excited. Similarly, it is preferable to have a good phase function, as discussed below, at low frequencies rather than high ones if a choice must be made.

Sometimes it is desirable to use a servomechanism as a filter to remove from the output motion certain bands of frequencies which are present in the input. A common filter removes high frequency noise from the input, and this then determines the band of frequency response.

3. The phase characteristics as in Figure 4 also carry information concerning the response of a system, since the phase lag may be considered as a time delay for each component. In general, the best response is obtained with a system which has the least phase lag in the transfer function, particularly at the lower frequencies.

One important phase characteristic to recognize is the case for which the phase lag is proportional to frequency. In this case, since each component will be delayed by a time,

$$t_0 = \frac{\text{phase lag in radians}}{\text{angular frequency in radians per second}}$$

all components will be delayed equally, and therefore the complete transient response will be delayed by a time t_0. A special case of this occurs when the servomechanism is following a constant velocity input at speed n; then the lag in following, after transients have died out, will be (nt_D) where t_D is the slope of the phase curve near zero frequency.

While it is often convenient to consider the phase functions in the interpretation of the transfer function, it is not absolutely necessary since there is a definite relationship between the phase and magnitude functions, as later discussion will show.

Electrical Torque Amplifier

While these criteria for good frequency response are quite qualitative in nature, they do provide a basis for the comparison of servomechanisms, and, more important, they provide a powerful aid in thinking about the various factors that can produce stability or instability in a system. This will be illustrated by an analysis of the torque amplifier of Figure 1, which is detailed in Figure 6.

The motor drive for the torque amplifier consists of a d-c motor and a controllable motor generator set. The servomotor has a constant field excitation and its armature voltage is supplied from the motor generator set. The generator is driven at constant speed, and its fields are in the output circuit of an amplifier which varies the excitation and therefore the generated voltage.

The amplifier is of the d-c type and has a push-pull output stage arranged so that the field flux is proportional to the output unbalance, which in turn is proportional to the input. Inductance in the field circuit prevents a sudden change in input voltage from producing a sudden change in field flux. This inductance and the consequent delay will have an effect on the stability of the system.

A network, often called a lead or derivative network, is included in the amplifier circuit for the purpose of stabilizing the system and improving its response, as later analysis will show.

The device for measuring the error between the input and output may, for example, be a reluctance type of bridge which provides an alternating voltage proportional to the error without placing an appreciable load on the input shaft. A direct voltage is obtained proportional to the error by means of a phase-sensitive detector, and a simple RC type filter is included to filter the detector output. The filter also will have an effect on the servo stability.

Analytical Derivation of the System Transfer Function

The most straightforward, but a tedious, procedure for the calculation of the transfer function is summarized as follows:

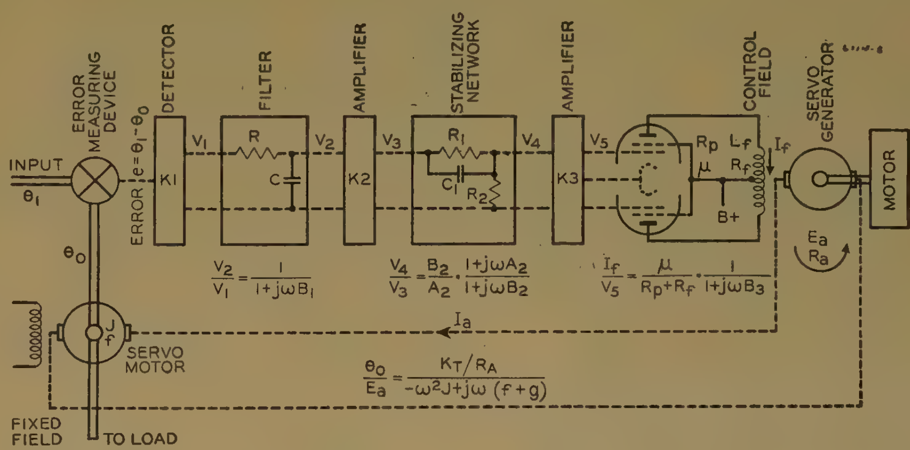


Figure 6. Torque amplifier

θ_o, θ_i = output and input positions in radians

$e=\theta_i-\theta_o$ = error or deviation in radians

V_1, V_2, V_3, V_4, V_5 = direct voltages as shown in figure

$K_1=\frac{V_1}{e}$, $K_2=\frac{V_3}{V_2}$, $K_3=\frac{V_5}{V_4}$ amplification or attenuation ratios

R, C = filter element values

$B_1=RC$ = filter time constant

R_1, R_2, C_1 = stabilizing network element values

$A_2=R_1C_1$ = stabilizing network time constant

$B_2=A_2\frac{R_2}{R_1+R_2}$ = stabilizing network delay time constant

R_p, μ = output tube plate resistance and amplification factor

L_f, R_f, ϕ_f = servo generator control field inductance, resistance, and flux

$B_3=\frac{L_f}{(R_p+R_f)}$ = servo generator control field time constant

E_a, I_a = armature circuit generated voltage and current

$T=K_TI_a$ = servomotor torque

J = servomotor plus load inertia

f = servomotor plus load viscous friction

g = effective viscous drag caused by motor back electromotive force

1. Set up the linear differential equations for the system.

2. Replace each derivative dx/dt by px and each integral $\int xdt$ by x/p just as in the transient solution and then solve the resulting set of simultaneous equations for the output-to-input ratio. For the system detailed in Figure 6 and with symbols given in the accompanying list, this expression comes out as

$$\frac{\theta_o}{\theta_i}=\frac{K(1+A_2p)}{\left\{\begin{array}{l}K+p(KA_2+f+g)+p^2[J+(f+g)(B_1+B_2+B_3)]+p^3[J(B_1+B_2+B_3)+(f+g)(B_1B_2+B_2B_3+B_3B_1)]+p^4[J(B_1B_2+B_2B_3+B_3B_1)+(f+g)(B_1B_2B_3)]+p^5(JB_1B_2B_3)\end{array}\right\}} \quad (1)$$

The denominator of this expression becomes the characteristic equation for the system. This is a fifth order equation which only can be solved numerically, and it illustrates the intricate interdependence of parameters which result from this approach. Since it is to serve only as an illustration and not for calculation, the detailed derivation is not given.

3. Substitute $p=j\omega$. This results in the equation

$$\frac{\theta_o}{\theta_i}=\frac{K(1+j\omega A_2)}{\left\{\begin{array}{l}K-\omega^2[J+(f+g)(B_1+B_2+\\ B_3)]+\omega^4[J(B_1B_2+B_2B_3+\\ B_3B_1)+(f+g)(B_1B_2B_3)]\end{array}\right\}+j\omega\left\{\begin{array}{l}(KA_2+f+g)-\omega^2[J(B_1+\\ B_2+B_3)+(f+g)(B_1B_2+\\ B_2B_3+B_3B_1)]+\omega^4JB_1B_2B_3\end{array}\right\}} \quad (2)$$

This expression is known as the system response transfer function, and from it the magnitude and phase angle may be determined by the usual manipulations with complex numbers. Thus

$$\frac{\theta_o}{\theta_i}=\frac{R_N+jX_N}{R_D+jX_D} \quad (3)$$

where the R's and X's are all real functions of the angular frequency $\omega=2\pi f$.

The magnitude of θ_o/θ_i is

$$M=\frac{\sqrt{R_N^2+X_N^2}}{\sqrt{R_D^2+X_D^2}} \quad (4a)$$

The phase angle of θ_o/θ_i is

$$\phi=\tan^{-1}\left(\frac{X_N}{R_N}\right)-\tan^{-1}\left(\frac{X_D}{R_D}\right) \quad (4b)$$

M and ϕ may be calculated by equations 4a–4b for particular parameter values to obtain plots such as in Figure 4. This analysis has eliminated the necessity for the solution of the characteristic equation, but the parameters still are intermixed so completely that interpretation is difficult.

The Loop Transfer Function

A second and simpler approach to the system transfer function is through the use of an auxiliary transfer function which leaves most of the system parameters independent. The type of automatic control under discussion is based upon the use of an error signal which is the difference between the input and output motions,

$$e=\theta_i-\theta_o=\text{error} \quad (5)$$

This expression yields

$$\frac{\theta_o}{\theta_i}=\frac{\left(\frac{\theta_o}{e}\right)}{1+\left(\frac{\theta_o}{e}\right)} \quad (6)$$

Equation 6 is of basic importance, since it states that the complex system transfer function may be expressed in terms of an auxiliary complex transfer function θ_o/e, which may be called the loop transfer function. It will be shown that the loop transfer function is simpler to derive and interpret than the system function, and a geometrical evaluation of equation 6 can be used. The term "loop transfer function" is borrowed from feed-back amplifier terminology,[3] and the similarity of θ_o/e to the often quoted $\mu\beta$ characteristic is apparent.

Reference to Figure 6 indicates that between the output θ_o and the error e there is a series of separate component circuits each with its own transfer function. Physically it is understandable that the magnitude of the resultant transfer function θ_o/e should be equal to the product of the magnitudes of the component transfer functions. Similarly, the resultant phase angle should be the sum of the component phase angles. This method of combination is also a property of the multiplication of complex numbers, so that the following equation may be written:

$$\frac{\theta_o}{e}=\frac{V_1}{e}\cdot\frac{V_2}{V_1}\cdot\frac{V_3}{V_2}\cdot\frac{V_4}{V_3}\cdot\frac{V_5}{V_4}\cdot\frac{I_f}{V_5}\cdot\frac{\phi_f}{I_f}\cdot\frac{E_a}{\phi_f}\cdot\frac{I_a}{E_a}\cdot\frac{T}{I_a}\cdot\frac{\theta_o}{T} \quad (7)$$

in which each transfer function is a complex number.

The next step is to calculate all the individual transfer functions for substitution in equation 7. This step requires only simple electrical circuit theory.

For the filter,

$$\frac{V_1}{e}\cdot\frac{V_2}{V_1}=K_1\cdot\frac{1}{1+j\omega RC}=K_1\cdot\frac{1}{1+j\omega B_1} \quad (8)$$

For the stabilizing network,

$$\frac{V_3}{V_2}\cdot\frac{V_4}{V_3}=K_2\frac{R_2}{R_1+R_2}\times\frac{1+j\omega R_1C_1}{1+j\omega R_1C_1\cdot\frac{R_2}{R_1+R_2}}=K_2\frac{B_2}{A_2}\cdot\frac{1+j\omega A_2}{1+j\omega B_2} \quad (9)$$

For the generator field current,

$$\frac{V_5}{V_4}\cdot\frac{I_f}{V_5}=K_3\cdot\frac{\mu}{R_p+R_f+j\omega L_f}=K_3\cdot\frac{\mu}{R_p+R_f}\cdot\frac{1}{1+j\omega\frac{L_f}{R_p+R_f}}=\frac{K_3\mu}{R_p+R_f}\cdot\frac{1}{1+j\omega B_3} \quad (10)$$

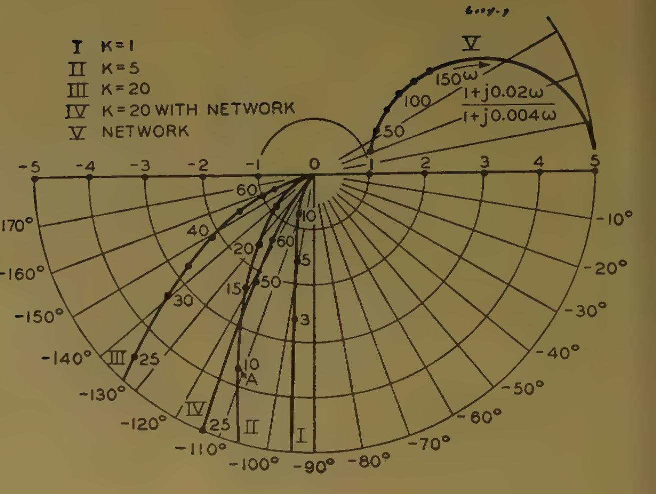
For the generated voltage,

$$\frac{\phi_f}{I_f}\cdot\frac{E_a}{\phi_f}=K_4 \quad (11)$$

a constant which depends upon the generator design.

The terms of equation 7 between the generated voltage and the output motion cannot be separated into a series of individual transfer functions, since there is an

Figure 7. Loop transfer functions



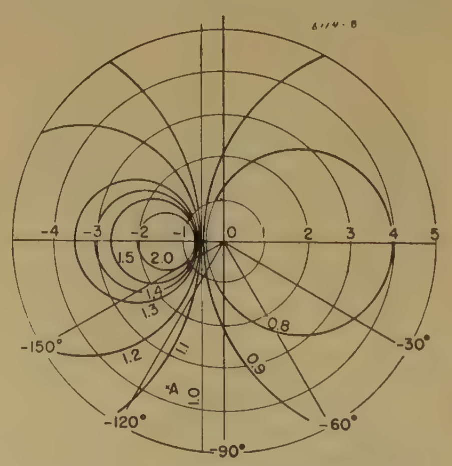


Figure 8. Loci of constant magnitude M on the θ_o/e complex plane

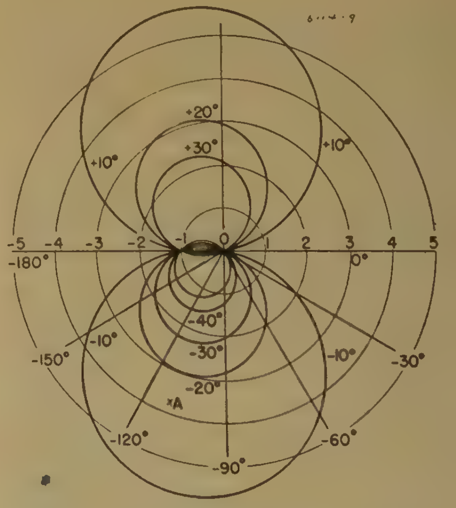


Figure 9. Loci of constant phase ϕ for θ_o/θ_i on the θ_o/e complex plane

interaction between the output motion and the armature current caused by the back electromotive force of the motor. The following three equations specify this interaction:

$$I_aR_a+K_m\frac{d\theta_o}{dt}=E_a$$

$$T=\text{torque}=K_TI_a$$

$$T=J\frac{d^2\theta_o}{dt^2}+f\frac{d\theta_o}{dt}$$

By setting $\frac{d}{dt}=j\omega$ and $\frac{d^2}{dt^2}=-\omega^2$ and combining the three equations:

$$\frac{\theta_o}{E_a}=\frac{\frac{K_T}{R_a}}{-\omega^2J+j\omega(f+g)} \qquad (12)$$

where

$$g=\frac{K_TK_M}{R_a}$$

Finally, for the loop transfer function, the combination of equations 8–12 into equation 7 yields

$$\frac{\theta_o}{e}=K\cdot\frac{1+j\omega A_2}{\left\{\begin{matrix}(1+j\omega B_1)\cdot(1+j\omega B_2)\times\\(1+j\omega B_3)\cdot(-\omega^2J+j\omega(f+g))\end{matrix}\right\}} \qquad (13)$$

where the A's and B's all have the dimensions of time, and the K lumps together all of the gain constants in the loop

$$K=K_1K_2\frac{B_2}{A_2}\cdot\frac{K_3\mu}{R_p+R_f}\cdot K_4\cdot\frac{K_T}{R_a}$$

Relation Between the Loop and System Transfer Functions

When equation 13 is introduced into equation 6, the system transfer function of equation 2 results. For the graphical interpretation of the relationship 6 between the loop and system transfer functions, it is desirable to plot the loop transfer function in a different manner, on polar co-ordinates. At any one frequency the transfer function has a magnitude and an angle which entirely specify a point on polar co-ordinates. As the frequency varies, the point will move and the resultant locus of points is the plot of the transfer function. Several such loci are shown in Figure 7 with the frequencies shown as numbers along each curve.

On the polar co-ordinate system on which these loop transfer functions are plotted, there must be a locus of points which all result in the same magnitude of system transfer function. These loci are the circles shown in Figure 8 for several different magnitudes. If any frequency point of θ_o/e falls on one of these circles, then the magnitude of the system transfer is determined by the value of M for that curve. Similarly, there are a series of equal phase angle curves ϕ as shown in Figure 9.

The fact that these loci are circles is based upon the representation of θ_o/e as a vector from the origin and $1+(\theta_o/e)$ as a vector to the same point from -1. Then $M=|\theta_o/\theta_i|$, according to equation 6, is the ratio of these two distances, and by a geometrical theorem the loci of constant M are circles of radii $M(1-M^2)^{-1}$ at centers $M^2(1-M^2)^{-1}$. Similarly, the angle ϕ is the difference between the angles of the two vectors, and the loci of constant ϕ are circles of radii $^1/_2\,|\text{cosec}\,\phi|$ at centers $-^1/_2+j^1/_2\cot\phi$.

The similarly numbered curves of Figure 7 and those of Figures 4 and 5 correspond to the same system. As an example of this method of calculation, consider the point A at $\omega=10$ for the loop transfer function II in Figure 7. This point falls on Figures 8 and 9 as shown and corresponds to $M=1.08$ and $\phi=-17°$. These values then appear on Figure 4 for the same frequency. With sufficient experience with this sort of calculation, it is no longer necessary always to plot the system transfer functions, and the engineer easily can interpret the characteristics of a system directly from the loop transfer functions of Figure 7.

If the loop transfer function falls in the neighborhood of the (-1) point, the resonance of the system becomes very high. Loci near the negative imaginary axis will not have resonance peaks. Also, for all large values of θ_o/e, the magnitude of θ_o/θ_i, will be near unity and the phase angle will be near zero.

The Torque Amplifier Loop Transfer Function

To obtain an understanding of the effect of each component network on a system, it is often desirable to study first a system with only the most important components and then gradually add the less important components. For the torque amplifier system of Figure 6, all the delays and the stabilizing network may be neglected at first, and the simplified loop transfer function

$$\frac{\theta_o}{e}=\frac{K}{-\omega^2J+j\omega(f+g)} \qquad (13)$$

may be studied.

Typical parameter values for a 0.1 horsepower system are

$J=0.005$-inch-ounces torque for one radian per second² acceleration of the motor.

$(f+g)=0.125$-inch-ounces torque for one radian per second velocity of the motor.

The value of K is determined by the amplification of the system. Experimentally, if the motor shaft is displaced one radian from zero error, the torque required to accomplish this is numerically equal to K. Curves I, II, and III are drawn for $K=1$, 5, and 20 inch ounces per radian respectively.

Since K is merely an amplification factor, the loop transfer functions I, II, and III of Figure 7 only are expanded or contracted versions of each other. For $K=1$, no resonance peak occurs, since the locus does not go near the point -1, but on the other hand the response drops off at a very low frequency. $K=5$ produces a system with a wider frequency response, but a resonance of 1.38 occurs at $\omega=25$ radians per second. $K=20$ has a very wide frequency response, but it also has an extremely high resonance peak.

From the standpoint of resonance, the system with $K=1$ is best, but its actual performance is poor, since the low frequency response results in a very sluggish system. Moreover, since it requires so much error to produce a torque, any load on the motor will produce a large error. For example, in order to drive a 10-inch-ounces dry friction load, a motor error of ten radians would be required at $K=1$-inch-ounce per radian. $K=5$ results in faster and tighter response, but even this is usually not enough for a high performance system. $K=20$ is much more desirable from every standpoint except that of the high resonance peak.

The design problem usually resolves into the necessity for providing some adequate means for reducing the resonance peak of the system with $K=20$. The network shown between V_4 and V_3 of Figure 6 is a common one used for this purpose. Its transfer function is

$$\frac{R_2\cdot(1+j\omega R_1C_1)}{(R_1+R_2)\left(1+j\omega R_1C_1\cdot\frac{R_2}{R_1+R_2}\right)}=\frac{B_2}{A_2}\cdot\frac{1+j\omega A_2}{1+j\omega B_2}$$

and its polar plot is shown in Figure 7 for $R_2/(R_1+R_2)=0.2$, and $R_1C_1=0.02$ second with K adjusted by changing other factors to take care of the attenuation of this circuit at $\omega=0$ which is 0.2

When this network is inserted in the system, the two complex transfer functions of the unstabilized system and the stabilizing network are multiplied together. Thus, at each frequency the two magnitudes are multiplied and the two phase angles are added so that the network shifts the loop transfer function away from the high resonance region near the -1 point. For the above values and $K=20$, the resulting loop transfer func-

tion is shown as IV in Figure 7, and the system transfer function is shown as IV in Figure 4.

The introduction of the stabilizing network has reduced the resonant peak to 1.14 at 50 radians per second and also has broadened the frequency response and reduced the phase lag.

An interesting question which illustrates the power of this method is the determination of the proper value for the time constant R_1C_1. The form of the locus which is a circle is independent of R_1C_1, but the position of each frequency point is determined by R_1C_1. For example, if a low time constant of say only 0.002 second had been chosen, the stabilizing network would not be able to introduce an appreciable positive phase angle until the very high frequency region after the loop transfer function already had passed through the highly resonant region. Conversely, if R_1C_1 is too large, say 0.2 second, the large positive phase angles would occur at very low frequencies and the phase would have returned to near zero at the frequencies for which the loop transfer function passes near the point -1. This relationship is shown in Figure 10 by an exaggerated sketch.

The Effect of Delays

With the addition and proper design of the stabilizing network, the torque amplifier has been given quite satisfactory characteristics, but certain delays which initially were assumed to be of secondary importance still are to be investigated. The transfer function for a single delay has been given in equation 10, and it is well known that its locus is a circle with negative phase angles, as shown by curve E of Figure 11. Since such a delay adds a negative phase angle into the loop transfer function it will shift the locus toward the (-1) point and cause an increase in the resonance peak. A delay of this sort is to be considered important or not depending upon whether or not it introduces an appreciable phase shift at the frequencies near the resonant point.

Figure 11, curve A, shows the locus of the stabilized system without delays. Curve B shows the same locus as it is shifted by a 0.004-second delay which might occur with a pentode tube driving the field circuit. In this case the shift increases the resonance to 1.5 from its previous value of 1.14. Curve C shows the effect of a single 0.02-second delay as might occur with a triode driving tube. This delay is quite appreciable and the resonance has risen to almost four.

To illustrate another feature, curve F is the locus of the transfer function for two 0.01-second delays and curve D shows their effect on the system. These two delays might occur in the filter network and in the field circuit, respectively. They have the transfer function $(1+j.01\omega)^{-2}$ and are shown to have a greater effect than that of a single 0.02-second delay.

Curve D of Figure 11 represents a system that is worse than just resonant; it is unstable, and oscillations increase in amplitude with time, rather than die out. Any locus for which the vector from the (-1) point to the locus rotates through a resultant negative angle as ω goes from zero to infinity will be unstable. Thus, curve C could represent also an unstable system if the gain were increased until the locus circled the (-1) point. For a fully rigorous statement and proof of this stability criterion, usually called the Nyquist criterion, reference should be made to the writings of Nyquist,[5] Bode,[7] and MacColl.[8]

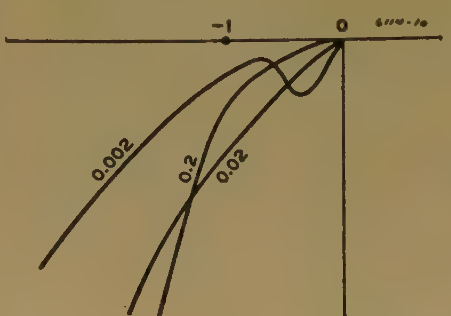


Figure 10. Effect of the stabilizing network time constant on θ_0/e

In the event that these delays are appreciable, it is necessary to reduce the loop gain or add other suitable phase correcting networks, but since this is only an illustrative example, the design will not be completed here.

Inverse Transfer Functions

Equation 6 may be placed in a form which still further simplifies the calculating procedure in many cases. The reciprocal of equation 6 is

$$\frac{\theta_i}{\theta_o}=\frac{e}{\theta_o}+1 \tag{14}$$

In this equation θ_i/θ_o and e/θ_o represent an input-to-output ratio rather than an output-to-input ratio and therefore are called the inverse system transfer function and the inverse loop transfer function, respectively. The graphical solution of equation 6 is particularly simple since, if e/θ_o is plotted on the complex plane as before, the same curve may be used for θ_i/θ_o with a new origin at the (-1) point. Loci of constant magnitude and constant phase for θ_i/θ_o are the concentric circles and intersecting lines of the polar co-ordinate system at the (-1) point. Thus $M^{-1}=|\theta_i/\theta_o|$ is the distance from the (-1) point to the locus, and $(-\phi)$ is the angle of the vector from the (-1) point to the locus.

The Calculus of Transfer Functions

When several components or complete systems are placed in cascade, their transfer functions combine as

$$Me^{j\phi}=M_1e^{j\phi_1}M_2e^{j\phi_2}=M_1M_2e^{j(\phi_1+\phi_2)} \tag{15}$$

This relation was used in equation 7.

Similarly the inverse transfer functions which have the reciprocal magnitude and the negative phase angle of the transfer functions combine as

$$M^{-1}e^{j(-\phi)}=M_1^{-1}M_2^{-1}e^{j(-\phi_1-\phi_2)} \tag{16}$$

When one component feeds back from the output of another component as shown in Figure 12, the resultant transfer function is

$$Me^{j\phi}=\frac{M_1e^{j\phi_1}}{1+M_1M_2e^{j(\phi_1+\phi_2)}} \tag{17}$$

while the inverse transfer functions combine as,

$$M^{-1}e^{j(-\phi)}=M_1^{-1}e^{j(-\phi_1)}+M_2e^{j\phi_2} \tag{18}$$

Since this is only a vector addition, its graphical construction is very simple and provides one of the advantages of the inverse transfer function approach.[6] This is particularly so in the study of regulators and process controls. Another advantage is that it places the loci near the origin for the low frequencies which are of interest.

Regulator Principles

For the torque amplifier example, only its response to an input motion has been considered, and it was desirable for the system to have a wide frequency response. If the torque load on the servomotor should vary, however, it would be desirable to have a minimum response to these variations. That is, if torque variations are considered as an input, the output θ_o should be filtered as completely from these variations as is possible. In this sense the torque amplifier is acting as a regulator, and the use of equation 18 with inverse transfer functions becomes advantageous.

If a torque T_L is applied at the load shaft with the armature circuit connected

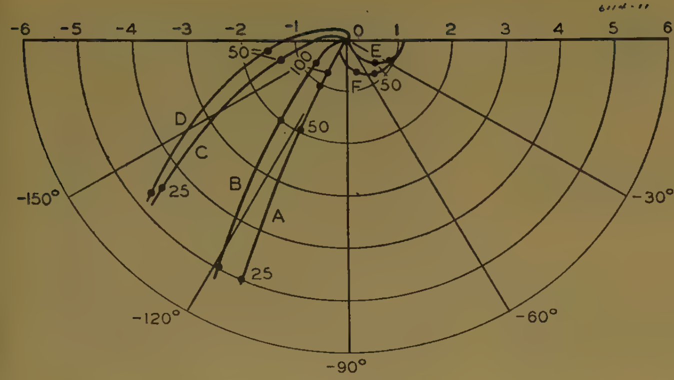
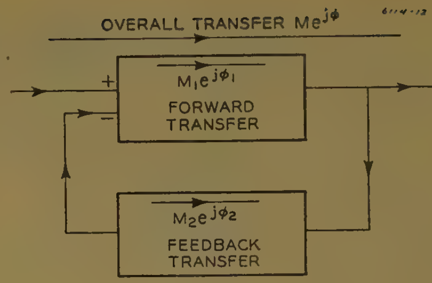


Figure 11. Loop transfer function for stabilized system with delay

A—Basic system with no delays
B—A 0.004-second delay
C—A 0.02-second delay
D—Two 0.01-second delays

Figure 12 (right). Feedback over a component or system



but with the amplifier off, the resultant output will be calculated as

$$T_L = Jp^2\theta_o + fp\theta_o - T$$

$$T = K_T I_a = -\frac{K_T K_M}{R_a} p\theta_o = -gp\theta_o$$

$$\frac{T_L}{\theta_o} = p^2J + p(f+g) \tag{19}$$

T_L/θ_o is an inverse transfer function and may be identified with $M_1^{-1}e^{-j\phi_1}$ of equation 18.

The return transfer function $M_2e^{j\phi_2}$, which is the ratio of the torque produced by a motion of the output from the zero error position, is (with $B_1 = B_3 = 0$)

$$M_2e^{j\phi_2} = K \cdot \frac{1+j\omega A_2}{1+j\omega B_2} \tag{20}$$

Combination of equations 19 and 20 into equation 18 yields

$$\frac{T_L}{\theta_o} = p^2J + p(f+g) + K \cdot \frac{1+j\omega A_2}{1+j\omega B_2} \tag{21}$$

for the system with the amplifier on.

T_L/θ_o of equation 19 represents the inverse transfer function for a disturbing torque with the amplifier off, and as such it is a measure of the self-regulating properties of the load. It is plotted in Figure 13, curve *A*, for the values $J = 0.005$-inch-ounce per radian per second^2 and $(f+g) = 0.125$-inch-ounce per radian per second. Thus at $\omega = 10$, T_L/θ_o is only slightly greater than one and only 1-inch-ounce torque is required to move the load with an amplitude of one radian at this frequency.

T_L/θ_o of equation 21 is the same transfer function with the control loop closed. Curve *B* in Figure 13 is the plot for this function, with $K = 20$, $A_2 = 0.02$, and $B_2 = 0.004$. T_L/θ_o is greater than 20 at all frequencies for this system, and this is a measure of how well the torque amplifier is regulated against load changes.

The equivalent condition to a high resonance peak would occur if the locus came relatively close to the origin for certain frequencies. Then disturbances at these frequencies would be regulated less well than disturbances at other frequencies.

Phase and Magnitude Relationships

In the foregoing discussion the phase and magnitude of the transfer function have been treated as more or less independent quantities. Actually since they are derived from the same function of a complex variable, there must be a relation between the two which definitely limits the variety of polar plots which are available in physical systems. Bode[7] has developed several such relations which are useful in feed-back amplifier design and consequently in automatic control design. The dependence of phase upon magnitude for these linear stable systems can be written only in an implicit integral form, but its essence may be stated simply as follows.

In physically realizable stable systems of the kind we are discussing, there is a tendency for the phase angle of a transfer function to depend upon the rate of increase or decrease of the magnitude of the function with frequency. If the magnitude decreases as $1/\omega^k$ over a region of frequency, then the phase will tend toward an angle of $-k\pi/2$ radians. The simple delay $(1+jB\omega)^{-1}$ illustrates this, since at low frequencies the magnitude is slowly varying and the phase is near 0 degrees, while at high frequencies the magnitude decreases as $1/\omega$ and the phase approaches -90 degrees. The transition between two rates of change of magnitude results in a smooth variation in phase between the two values. All the transfer functions plotted illustrate this relation ship.

The important feature to note is that if the loop transfer function magnitude decreases as $1/\omega^2$ or faster for a region near unity, the phase angle will be near $-180°$ or more and the system will be highly resonant. A good loop transfer function should not decrease in magnitude any faster than, say, $1/\omega^{1.5}$ until the magnitude is below about 1. Then the resonance peak will be no greater than 1.3. The stabilizing network discussed in the foregoing is useful, since it has a rising gain characteristic over a region near unity for which the loop magnitude is decreasing too fast. MacColl[8] has demonstrated this approach in detail.

Experimental Techniques

Experimentally many methods for testing a whole system for frequency response or for obtaining the transfer function of individual components will suggest themselves. A sinusoidal disturbance is placed

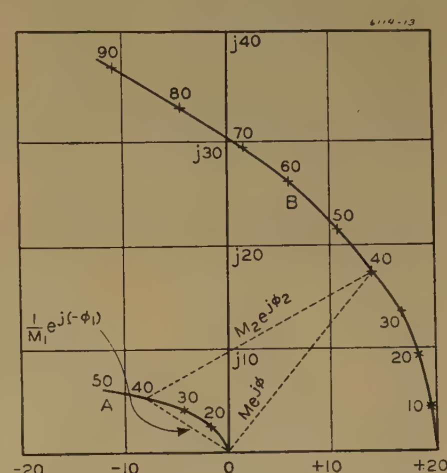


Figure 13. The open and closed loop inverse transfer functions for a disturbing torque

at the input of the component or system, and the relative phase and magnitude of the output is measured. One important advantage of this type of analysis is that the transfer function for some critical component which is difficult to analyze may be determined experimentally and

Factors Affecting the Range of Radar Sets

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IT is now well known that most of the radar sets used in the past war function by transmitting a pulse and measuring the time necessary for the signal to reach the target and the echo to return. Of particular interest is the problem of finding how much power must be transmitted so that the echo will be of sufficient strength to be detected satisfactorily by the radar receiver. When dealing with microwave equipment and the consequent line-of-sight transmission paths, the problem can be solved by geometric means.

Consider a point source which is radiating energy uniformly in all directions.* The rate at which energy passes through unit area at a distance r from the source is $P_t/4\pi r^2$ where P_t is the power output of the transmitter. This must be true because the area of the sphere of radius r, through whose surface all the energy passes, is $4\pi r^2$.

A radar transmitter which radiates equally in all directions is difficult to obtain. However, a very common short wave antenna is the dipole which has a radiation pattern similar to that shown in Figure 1. This is a polar diagram of the field strength through a plane containing the dipole. The flow of energy in directions A is roughly 3/2 that of the ideal point source and hence the power gain of the dipole is said to be 3/2. Thus the energy flow through unit area at a distance r from the dipole, in the directions A, is $3P_t/8\pi r^2$.

* This discussion is adapted in part from a series of lectures given by W. W. Hansen at the Radiation Laboratory, Massachusetts Institute of Technology, in 1940 and 1941.

Viewed as a receiving system, the so-called effective cross section of a dipole[1] is a function of the wave length, λ, and is approximately equal to $3\lambda^2/8\pi$. This is numerically equal to the cross section of a disk which, if placed normal to the incident radiation, will absorb an equivalent amount of energy. Thus if two similar dipoles are placed a distance r apart, the received power, P_r, in terms of the transmitted power, will be the product of the energy density at the receiving dipole and the dipole cross section,

$$P_r = \frac{3P_t}{8\pi r^2} \times \frac{3\lambda^2}{8\pi} = P_t\left(\frac{3\lambda}{8\pi r}\right)^2 \qquad (1)$$

At microwave frequencies, it is convenient to use a parabolic reflector, with the dipole at its focus to narrow the radiated energy into a beam. Figure 2 is a sketch of this arrangement showing the dipole at the focus of a paraboloid. The effectiveness of this reflector is expressed in terms of the energy density along the axis of the paraboloid compared with the energy density at the same point attributable to the dipole alone, or, remembering the reciprocity theorem, in terms of the increase in the magnitude of a received signal resulting from the addition of the reflector. Using the latter line of reasoning, a first approximation to the power gain, G, of the reflector can be obtained by dividing the cross section area of the paraboloid by the effective cross section of the dipole,

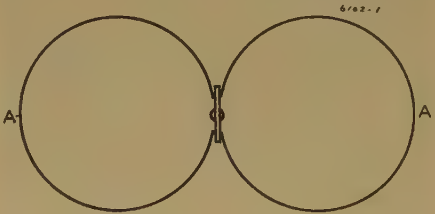

Figure 1. Field pattern of dipole

$$G = \frac{\dfrac{\pi D^2}{4}}{\dfrac{3\lambda^2}{8\pi}} = \frac{2}{3}\left(\frac{\pi D}{\lambda}\right)^2 \qquad (2)$$

where D is the diameter of the paraboloid and λ the operating wave length. The beam width θ between half-power points in terms of reflector size and wave length is given roughly by

$$\theta = \lambda/D \text{ radians} \qquad (3)$$

Referring to equation 1, if paraboloids are used behind each dipole, then the received power is increased by the product of the gains of these reflectors. Assuming paraboloids of equal size, the expression for the received power P_r becomes

$$P_r = P_t\left(\frac{3\lambda}{8\pi r}\right)^2 \times G^2$$

$$P_r = P_t\left(\frac{\pi D^2}{4\lambda}\right)^2 \times \frac{1}{r^2} \qquad (4)$$

As might be expected, this equation shows that the received power is inversely proportional to the square of the separation.

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placed along with other transfer functions which are determined analytically.

In some cases it may be desirable to work with the magnitude of the transfer function alone because of difficulty in making phase measurements. Then the phase-magnitude relationship is useful.

For experimental verification of the methods discussed, it is important that the system remain linear; in particular, the applied disturbances should not be large enough to overload the amplifiers or saturate the motor. In view of this limitation, this analysis yields necessary conditions for good performance but not always sufficient conditions.

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It is the fundamental relation describing point-to-point operation, such as microwave radiotelephone service where a line-of-sight path is involved.

For the radar case, however, it is necessary to find the energy in the echo. If the target is an airplane, the energy scattered is a function of the size of the airplane, materials, wave length, aspect, polarization of the incident wave, and so forth. An effective cross section A_t can be defined as the area intercepting that amount of power which, if scattered equally in all directions, will give an echo equal to that from the target.

With this assumption, the energy density at a distance r from the target is

$$P_s = P_t \times \frac{A_t}{4\pi r^2} \tag{5}$$

where P_t is the incident energy density and P_s the density of the scattered energy at a distance r from the target.

In a microwave set the same antenna system is used for transmitting and receiving; hence the received energy is the product of equations 4 and 5, since equation 4 represents the energy which would be received if it were reflected without loss of any sort and equation 5 is the loss factor of the target.

$$P_r = P_t\left(\frac{\pi D^2}{4\lambda}\right)^2 \times \frac{A_t}{4\pi} \times \frac{1}{r^4} \tag{6}$$

Will this energy be sufficient to operate the receiver? While it is possible to get almost infinite gain in a receiver, the sensitivity, that is, the smallest signal that the receiver can utilize, is determined by the noise in the receiver input.

The theoretical lower limit of this noise level is fixed by the thermal-agitation-noise (Johnson noise) in the resistive component of the input.[2,3] This thermal-agitation-noise voltage is $\sqrt{4kTBR}$ volts. Remembering that the maximum power which can be drawn from a generator of electromotive force E and internal resistance R is $E^2/4R$, the maximum noise power available to the receiver is kTB watts. Here k is Boltzman's constant, 1.37×10^{-23} Joule per degree absolute, T about 290 degrees absolute; and B the effective noise band width of the receiver in cycles per second, which is roughly equal to the band width of the receiver between half-power points. For satisfactory pulse reception, the band width should be inversely proportional to the length of the pulse. For a one-microsecond pulse, this is 10^6 cycles. The ratio of the equivalent noise level at the receiver input to the thermal-agitation-noise is defined as the "noise figure," symbol F,

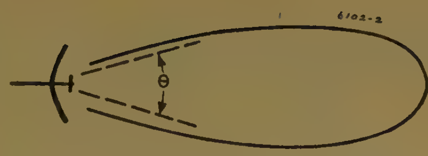


Figure 2. Field with parabolic reflector

and in a microwave receiver may be as low as four or five.

An expression for the maximum range of a radar set can now be written. If it is assumed that a received signal-to-noise ratio of one is acceptable, then the value of the equivalent noise level can be substituted for P_r in equation 6 giving

$$kTBF = P_t\left(\frac{\pi D^2}{4\lambda}\right)^2 \times \frac{A_t}{4\pi} \times \frac{1}{r^4} \tag{7}$$

Solving this for the range r gives the relation

$$r = \sqrt[4]{P_t\left(\frac{\pi D^2}{4\lambda}\right)^2 \times \frac{A_t}{4\pi} \times \frac{1}{kTBF}} \tag{8}$$

To illustrate by a numerical problem, the following is representative of a microwave search radar:

P_t, peak transmitted power, $=100$ kw
λ, wave length, $=10$ centimeters
Beam width $=3$ degrees, or,
D, diameter of reflector, $=190$ centimeters
A_t, effective cross section area of a medium bomber, $=2 \times 10^5$ square centimeters
B, band width, $=10^6$ cycles per second. (This permits a 1-microsecond pulse)
F, noise figure, $=10$

The maximum range is then roughly 150 land miles. The airplane could be tracked this far (in theory) if it flew at sufficient altitude to permit a line-of-sight path between it and the radar antenna.

Several interesting conclusions can be drawn from equation 8. The effective range varies at the fourth root of the transmitted power, that is, to double the range it is necessary to increase the peak power by a factor of 16. The range also varies as the fourth root of the effective cross section area of the target. For this reason a large bomber cannot be tracked to a much greater distance than a small fighter. Again, the range appears to be inversely proportional to the square root of the wave length, but this is not the whole story. As shown in equation 3, the beam width is directly proportional to the wave length, and if one endeavors to increase the range by decreasing the wave length with the consequent decrease in beam width, the beam speedily becomes too narrow for successful searching. In other words, it is possible for the beam to pass completely over the target between pulses. Actually it is desirable for at least three or four pulses to strike the target while the beam is sweeping over that point.

The range also appears to vary directly with the diameter of the reflector. But here again one runs into the problem of decreasing size of the beam and the consequent unsuccessful scanning. If the beam width is held constant, which means that the ratio of D to λ is fixed, then the range actually falls off with decrease in wave length, because the size of the reflector must be reduced in proportion to the reduction in λ. It should be added here that a paraboloid, such as was assumed in the foregoing numerical example, is not very practical for a long range search set because the beam is too narrow in the vertical direction. One way to overcome this difficulty is to use a truncated paraboloid which has a considerably wider beam in elevation than in azimuth. The gain, of course, of such a reflector is less than the complete paraboloid.

The range is also proportional to the fourth root of the peak power of a pulse. But increasing the peak power by shortening the pulse to keep the same average power also requires that the band width be increased; hence the noise level of the receiver will increase in proportion. If the peak power is increased without shortening the pulse, which means that the average power is increased, then, of course, the range also will increase, but slowly.

Finally, it is seen that the range is inversely proportional to the fourth root of the noise figure. In fact, as far as this characteristic is concerned, decreasing the noise figure is as effective as increasing the average power. This explains why such a tremendous amount of work has been done in the direction of reducing the noise figure of radar receivers. Reducing the noise figure by three decibels may not be as sensational as increasing the peak power from one megawatt to two megawatts, but the increase in range is just as great.

The recent experiments which resulted in obtaining radar echoes from the moon have been described in a recent publication.[4] After making certain corrections for the increase in gain of the radar antenna caused by reflections from the ground and so forth, it was found that the strength of the received echo agreed closely with the value predicted by calculation. A long-wave radar set was used (110-megacycle frequency and a pulse length, of 0.02 second or more). This long pulse allowed the over-all receiver band width to be reduced to 50 cycles per second with a consequent reduction in the thermal-agitation noise. A peak trans-

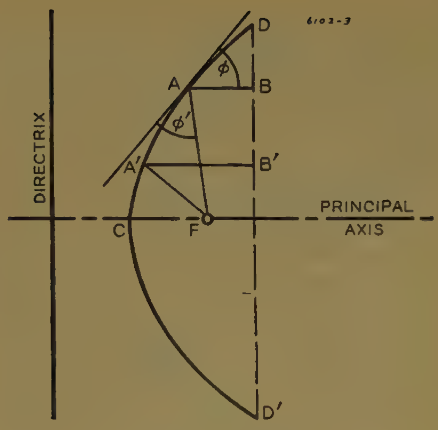


Figure 3. Parabola

mitted power of three kw gave a very adequate signal.

Earlier in this paper reference was made to a simple relation for the gain of a parabolic reflector (equation 2), and an idealized field pattern with only one lobe was shown in Figure 2. The true picture is somewhat more complicated. Because the diameter, or aperture, of the reflector is at the most a few dozen wave lengths, the field distribution is the result of diffraction of the radiation at the aperture. Hence the gain depends on other factors beside the size of the reflector and the wave length, and the field shows side lobes in addition to the main lobe.

A complete mathematical discussion of the action of the dipole and parabolic reflector is long and rather involved. However, a qualitative discussion will give a reasonably clear picture.

By definition a parabola is the locus of a point whose distance from a fixed point, the focus, is equal to its distance from a fixed line, the directrix (Figure 3). The parabola also has the property that the angle which any focal radius, such as FA, makes with a tangent to the parabola at the point of intersection is equal to the angle the tangent makes with a line through this point of intersection and parallel to the principal axis. These angles are shown as ϕ and ϕ' on the figure.

A geometric construction shows at once that the distance from the focus to any point on the parabola plus the distance from that point to a line parallel to the

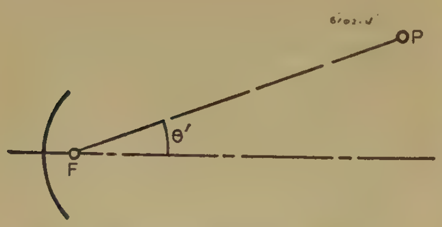
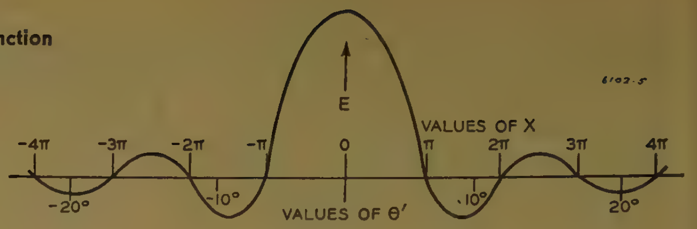


Figure 4. Location of point P

Figure 5. $\dfrac{\text{Sin } X}{X}$ function



directrix but situated on the other side of the vertex is a constant. Referring again to Figure 3, this means that $FA+AB=FA'+AB'$, and so forth.

If a source of radiation is placed at F and DCD' is a parabolic reflector, all the waves reflected from it will arrive at DD' in time phase. Therefore we can consider DD' (actually the disk equal to the aperture of the parabola) as a constant phase surface radiating to the right. Next this constant phase surface can be divided into elemental areas and the contribution of all these areas to the field at the point under consideration summed. In general, the distribution of the field intensity will have the form

$$E=(\text{constant})(1+\cos\theta')\,\frac{\sin X}{X} \tag{9}$$

where

$$X=\frac{\pi D}{\lambda}\sin\theta'$$

D = diameter of paraboloid DD'

θ' = angle between principal axis and line from focus to point under consideration (Figure 4)

Figure 5 is a plot of this function for D/λ roughly equal to ten, and Figure 6 is a polar plot of the field for a somewhat smaller reflector. The pattern of field intensity can be altered drastically by changing the focal length of the parabola, by distorting the reflector, and by varying the field intensity distribution across the aperture. The presence of direct forward radiation from the dipole feed also modifies the pattern. If the dipole is placed an odd number of quarter-wave lengths from the vertex of the parabolic reflector, then the forward radiation will reinforce the reflected wave (remembering that a wave is reflected with a 180-degree reversal in phase), and if placed an even number of quarter-wave lengths will tend to cancel it.

A typical design of a parabolic reflector calls for a ratio of focal length to diameter equal to 0.30 and also specifies that the product of the focal length and the depth shall equal one-fourth the diameter. Under these conditions, with a dipole feed, the gain is roughly that given in equation 2 and the beam width between half-power

Figure 6. Typical antenna pattern showing side lobes

points 1.2 λ/D radians. With a reflector of reasonable size, the side lobes are small enough to be unobjectionable.

The relations just discussed do not take into account any of the phenomena associated with refraction or reflection from external surfaces, or the fact that very short waves are absorbed by fog and rain.[5] For wave lengths much less than one inch, the latter problem is serious. Reflection can be both an aid and a hindrance. Properly situated ground radar search sets are subject to increases in received signal strength of as much as 12 decibels caused by ground reflections. On the other hand, when the target and the radar set are near the surface of the earth, it is entirely possible for the difference in path lengths of a direct and reflected echo (that is, the ray reflected from the surface of the ocean) to be some odd multiple of a half-wave length, and in this case the resultant echo field will be very nearly zero. This phenomenon has been encountered when searching for ships with shore-based radar sets.

Further work is being done on the measurement of the absorption by the atmosphere of wave lengths less than one inch. As might be expected, there are strong absorption bands due to oxygen, water vapor, and so forth. The results will have considerable bearing on future selection of wave lengths.

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The Preservative Treatment of Various Species for Poles and Crossarms

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THE heavy drain that the war has imposed on the forests of America, the backlog of pole requirements that has been built up over the war period, and the present labor shortage have made it impracticable to obtain a sufficient number of poles of the species formerly used.

Up to the present time, southern yellow pine and western red cedar have been the species most extensively used for poles. Coast Douglas fir and northern white cedar also are recognized as good pole woods, but poles of these species have not had the wide distribution nor the extensive use that southern yellow pine and western red cedar poles have had. In addition to the woods mentioned, treated lodgepole pine poles have been used to a limited extent in the Rocky Mountain region.

Cedar poles and in certain localities some of the less durable species have been used in past years without preservative treatment, but the present discussion will be confined to poles that are to be treated. Since the sapwood of all species has low resistance to decay, the economy of preservative treatment is recognized even for cedar poles that normally have a fairly thin sapwood and a naturally durable heartwood.

In order to meet some of the problems resulting from the current pole shortage, specifications have recently been prepared covering pole species to be used by the Rural Electrification Administration. The American Standards Association has prepared specifications designated as "American War Standards Specifications and Dimensions for Wood Poles." The specifications mentioned cover miscellaneous conifers that are to be given a preservative treatment and that heretofore have not been included among the recognized standard pole woods.

The proposed substitute woods have been grouped according to the allowable fiber stress in the following order:

Group I. Fiber stress 5,600 pounds per square inch

(a). Atlantic white cedar (*Chamaecyparis thyoides*)
(b). Spruce (*Picea*)—all species

Group II. Fiber stress 6,000 pounds per square inch

(a). Eastern white pine (*Pinus strobus*)
(b). Ponderosa pine (*Pinus ponderosa*)
(c). Sugar pine (*Pinus lambertiana*)
(d). Western white pine (*Pinus monticola*)

Group III. Fiber stress 6,600 pounds per square inch

(a). Jack pine (*Pinus banksiana*)
(b). Red (Norway) pine (*Pinus resinosa*)
(c). White fir (*Abies concolor*)

Group IV. Fiber stress 7,400 pounds per square inch

(a). Douglas fir other than coast type (*Pseudotsuga taxifolia*)
(b). Eastern hemlock (*Tsuga canadensis*)
(c). Western hemlock (*Tsuga heterophylla*)
(d). Eastern larch (Tamarack) (*Larix laricina*)
(e). Western larch (*Larix occidentalis*)

The foregoing fiber stresses were obtained from a study made by the Forest Products Laboratory and are based on the American Standards Association recommended stresses for lodgepole pine.

Wood Preservatives for Poles

Creosote

Prior to the war a large proportion of the poles used in the United States were treated with American Wood Preservers Association specification grade one coal tar creosote with a specified distillation residue of not over 20 to 25 per cent above 355 degrees centigrade.

The federal specification covering creosote for pole treatment limits the residue to 25 per cent. Shortages resulting from the war, however, made it difficult to obtain cresote with a low distillation residue, and it became necessary to increase the allowable residue in the federal specification to as much as 35 per cent. This wartime emergency provision has been discontinued. An important reason for specifying low residue cresote oils for pole treatment is to reduce the tendency to bleed.

Before the war a large amount of creosote was imported from Europe and Japan. Because this supply practically has been cut off and because the domestic production probably will not be able to meet the heavy postwar demands, a creosote shortage may exist for some time to come. This situation has forced pole users to consider the use of other preservatives less well known but that offer promising results. In extreme cases it even may be necessary to use treatments that are known to be less economical because of the shorter service life obtained. Such treatments may be less economical because the materials lack the preservative properties suitable for conditions under which poles are used or because of inadequate treatment. Nevertheless, there should be few cases where inferior and uneconomical treatment is necessary because of present shortages.

Chlorinated Phenols

During the past ten years, the chlorinated phenols have been receiving considerable attention as possible substitutes for creosote or as materials that can be used in mixtures with creosote.[1,2] Pentachlorophenol is the best known in this group and has been the most extensively used, but tetra-chlorophenol and 2-chlororthophenylphenol, in mixtures with pentachlorophenol, have also been given consideration. Pentachlorophenol has the advantage of a lower water solubility than the other two phenolic compounds. Tests show that these chemicals have a high degree of toxicity, and five per cent solutions (on a weight basis) are apparently suitable when solution absorptions needed to obtain the necessary penetrations are employed. A five per cent solution of penetachlorophenol (on a weight basis) commonly is recommended at present. This is for absorptions similar to those specified for pole treatment with creosote.

Both toxicity and permanence are highly important qualities of a good wood preservative, but only long-time service tests provide a satisfactory measure of permanence, and such tests are needed to determine fully the relative merits of the chlorinated phenols. The limited number of service tests that have been started during the past few years indicate that the chlorinated phenols have a promising future, but little information is available on the performance of these solutions when applied by the pressure process. Data thus far obtained apparently indicate that solutions of the chlorinated phenols made with the heavier petroleum

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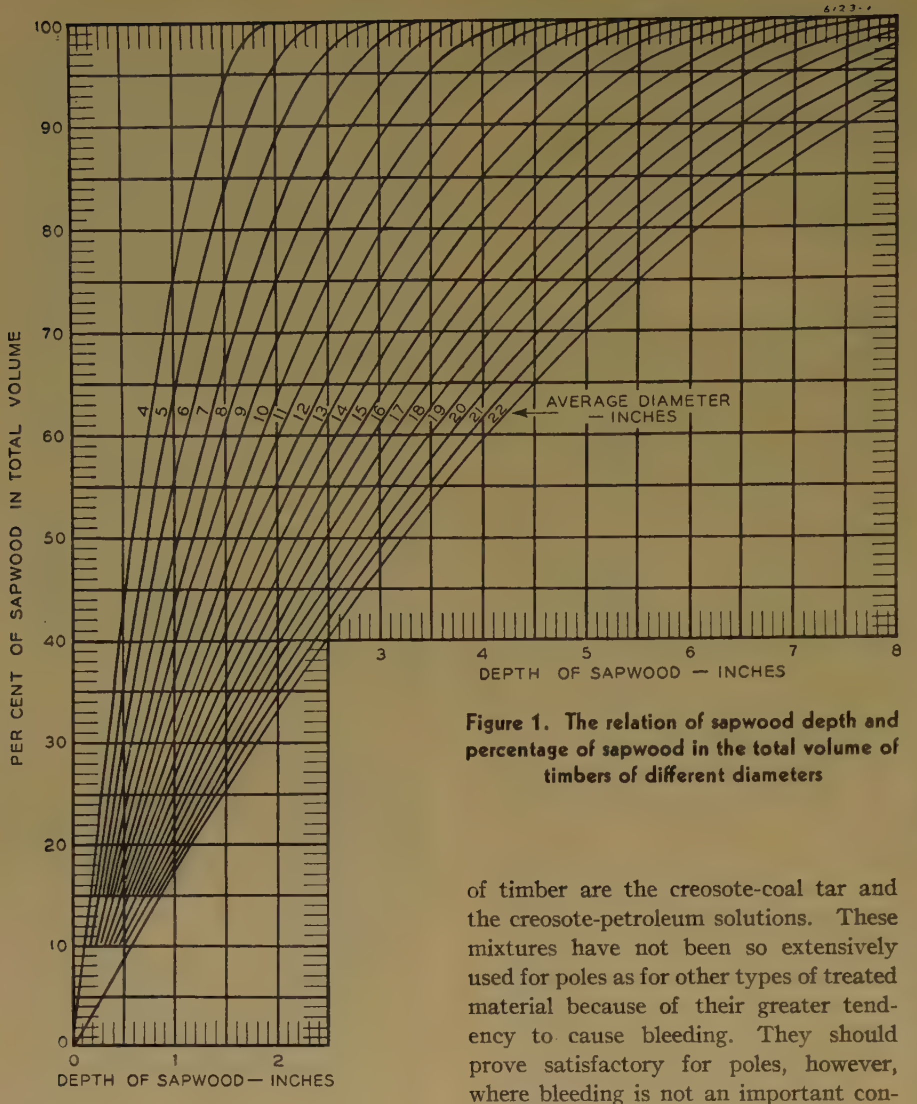

Figure 1. The relation of sapwood depth and percentage of sapwood in the total volume of timbers of different diameters

oils may give better protection than solutions in which the more volatile oils are used. On the other hand the more volatile oils penetrate the wood more readily when the solutions are applied by nonpressure methods. It has been found that some petroleum oils are more suitable than others both as solvents and from the standpoint of freedom from sludging.

The problem of sludging is also of importance in the use of creosote-petroleum mixtures. Experience has shown that petroleum oils with an asphaltic base are much less likely to give trouble than the oils with a paraffin base. Some pole users now are having poles treated with pentachlorophenol mixtures containing varying proportions of coal tar creosote. These mixtures also have an increased tendency to cause sludging unless a suitable petroleum oil is employed.

Creosote Solutions

Blended preservative oils that have been used successfully for various kinds of timber are the creosote-coal tar and the creosote-petroleum solutions. These mixtures have not been so extensively used for poles as for other types of treated material because of their greater tendency to cause bleeding. They should prove satisfactory for poles, however, where bleeding is not an important consideration.

Bleeding

The problem of bleeding is of particular interest when poles are used in urban line construction where the oily surface may cause damage to clothing and result in public complaint. Bleeding is naturally of less importance in rural lines, but in any case it is objectionable to linemen who must climb the poles when the line is installed and when repairs are required. The bleeding problem may be particularly aggravating when the poles have been treated shortly before they are placed in service. While this trouble becomes less acute after the poles have been in service for a time, in some installations bleeding may persist to a variable extent for a considerable period.

It must not be assumed that all poles treated with preservative oils bleed, but it is difficult to predict what poles, if any, will bleed and to what extent they will bleed.

Five factors that appear to have an important bearing on bleeding are

1. Type of preservative oil used.
2. Absorptions obtained.
3. Method of treatment.
4. Species.
5. Temperature conditions to which the wood is exposed.

Heavy absorptions usually cause more bleeding than light absorptions, heartwood will bleed more than sapwood, and species with the more resistant sapwoods will normally bleed more (with the same absorption) than species in which the sapwood is easily penetrated. Bleeding may occur either in summer or winter but is generally most severe in the hot summer months. Straight coal tar creosote usually gives less bleeding trouble than mixtures of coal tar creosote and petroleum, or creosote-coal tar solutions. Likewise, creosote with a high residue above 355 degrees centigrade usually causes more bleeding than low residue creosote oils.

Methods of Treatment

Pressure Treatment

The pressure process is the most effective method of treating timbers of all kinds, since it affords a means of controlling the pressure to any desired amount. By using different initial air pressures, a considerable variation in absorption can be obtained, thereby making it possible to get heavy or fairly light absorptions for a given depth of penetration. The principal objection to pressure treatment is that it requires relatively large and expensive equipment and is therefore not well adapted for the treatment of small quantities of timber nor for treating material that would require long distance transportation to the treating plant.

Nonpressure Treatment

Next to pressure treatment, the hot-and-cold bath method is the most effective. This treatment depends on atmospheric pressure utilized by first heating the wood in a heating medium and then cooling it in the preservative. When the wood is heated, the air contained within its cells is expanded, and part of that in the surface region of the timber is forced out of the lumina or air spaces. In the subsequent cooling bath, a partial vacuum is formed and more or less preservative is forced into the wood cells by atmospheric pressure, depending upon the permeability of the wood and upon the amount of air forced out by

expansion during the heating period. Very little preservative is absorbed during the hot bath, and it is obvious that the pressure forcing preservative into the wood during the cooling period cannot exceed atmospheric pressure. Generally the available pressure will be lower than atmospheric. For this reason the hot-and-cold bath treatment is more suitable for treating round timbers in which only the more easily treated sapwood needs to be penetrated. This method has been extensively used in the butt treatment of western red cedar poles and now is used to some extent in full length treatment of these poles. Lodgepole pine poles also have been butt treated by this method for use in the Rocky Mountain area.

Incising before treatment has been found very helpful when the sapwood is somewhat resistant to penetration, as in the case of cedar. When the sapwood is very resistant to penetration, as in species like the true firs and spruces, it is doubtful that even when the resistant sapwood is incised it will be possible to obtain satisfactory penetration without the use of pressure treatment.

The principal advantages of the hot-and-cold bath treatment over pressure treatment are

1. The cost of the treating equipment is small compared with the cost of pressure treating equipment.

2. Butt treatment or full length treatment can be used as desired.

3. It is more convenient to treat small quantities of timber by this method.

4. It is usually possible to make the treatments near where the timber is cut.

The principal disadvantages are

1. The pressure available for forcing the preservative into the wood is limited compared with the wide range of pressures available when the pressure process is used.

2. Considerably heavier absorptions usually are required for a given penetration than when the wood is treated by pressure methods because an initial air pressure cannot be applied as in the Rueping treatment.

3. This method of treatment is not well adapted for round timbers with unusually resistant sapwood nor for species that have a deep sapwood easily penetrated, since in species having a deep sapwood undesirably heavy retentions may be obtained or there may be incomplete sapwood penetration.

4. Evaporation of the lower boiling constituents of preservative oils used in the hot bath often causes an appreciable loss of preservative.

Absorption Specified for the Pressure Treatment of Poles

Most of the southern yellow pine and Douglas fir poles treated up to the present time have been treated by the empty-cell process with specified minimum net retentions of about eight pounds of creosote per cubic foot. Minimum absorptions specified for lodgepole pine poles have ranged from about four to eight pounds per cubic foot. The lower absorptions specified for the lodgepole pine have been possible partly because of the more shallow sapwood in this species. The sapwood of lodgepole pine commonly ranges from about one half to two inches in thickness compared with two to four inches in southern yellow pine and one to two inches in coast Douglas fir poles. Another reason for the use of lower absorptions in lodgepole pine is that poles of this species have been used mostly in the Rocky Mountain states where decay conditions are usually much less severe than those under which the southern yellow pine and coast Douglas fir poles are used.

Present American Wood Preservers Association specifications for the treatment of southern yellow pine poles permit a minimum net retention of six pounds per cubic foot provided the penetration specified for the 8-pound treatment is met and provided the poles are to be used under conditions that do not favor rapid decay. Various power companies have specified net retentions ranging from 10 to 16 pounds of either creosote or creosote-coal tar solutions in southern yellow pine poles.

The question of obsolescence is usually of more concern to telephone and telegraph companies than to power companies, and it is natural that they do not want to employ preservative treatments that may give a service life well beyond the time their pole lines would become obsolete. On the other hand, it obviously would be unwise and uneconomical to count on obsolescence and later find the service life was too short.

Sufficient service records are not yet available to show the minimum absorptions of different preservatives that can be expected to give satisfactory results for poles of different species when used under various climatic conditions. For this reason it would seem desirable to defer the extensive use of absorptions much lower than those commonly specified until the relative merits of the lower absorptions can be determined from suitable service tests. The following are some of the more important advantages of using absorptions that are known to be adequate as demonstrated by experience:

1. They furnish a better reserve against depletion by leaching and evaporation.

2. They insure better distribution and deeper penetration of the preservative.

3. When the preservative has a variable toxicity, the heavier absorptions serve as a safeguard against deficient toxicity.

4. They reduce the danger of insufficient absorption because of careless treatment.

5. They help protect against the possibility of inadequate treatment of the more resistant timbers in a charge or of those of the larger sizes when the charge contains timbers of different dimensions.

6. They reduce the danger of wide variations between the maximum and minimum penetrations obtained in the same or in different timbers.

The principal disadvantages of heavier absorptions are the somewhat higher initial cost of treatment and the possibility of objectionable bleeding when oily preservatives are used. The higher initial cost of treatment, however, does not necessarily mean that the annual charge will be higher over the period the timber is in service. Within reasonable limits the higher first cost may be much more than offset by the increased service life and consequent reduced cost of renewals.

There are a variety of factors that should be kept in mind in deciding upon the net retention that will prove most economical or most desirable because of other considerations.

The thickness of sapwood also will have a definite bearing on the absorptions that should be specified. Woods with deep sapwood, like that of ponderosa pine, red pine, and the southern pines, naturally have a greater proportion of wood that can be treated than poles of thin sapwood species, like the cedars, western larch, and mountain type Douglas fir. Figure 1 shows the proportion of sapwood in the total volume of poles of different average diameters ranging from 4 to 22 inches. These curves were computed from the relation that if t is the average depth of sapwood and D is the average diameter of the timber, the percentage of the sapwood P, based on the total volume, is given by

$$P=100\left[\frac{4t(D-t)}{D^2}\right] \qquad (1)$$

If P_t is the percentage of sapwood in the total volume when the average sapwood thickness is t inches, and P_T is the percentage of sapwood in the total volume when the average sapwood thickness is T inches, then

$$\frac{P_t}{P_T}=\frac{t}{T}\left[\frac{D-t}{D-T}\right] \qquad (2)$$

For example, assume $D = 10$ inches, $t = 0.75$ inch, and $T = 3$ inches. Then the ratio

$$\frac{P_t}{P_T}=\frac{0.75}{3}\left[\frac{10-0.75}{10-3}\right]=0.33$$

In other words, a 10-inch-diameter pole with a sapwood thickness of 0.75 inch would have about 33 per cent as much sapwood as a pole timber of the same diameter with a sapwood thickness of three inches. Equation 1 or Figure 1 shows that the percentage of sapwood in the total volume of a timber 10 inches in diameter with a sapwood thickness of 0.75 inch is about 27.7, while the same diameter timber with an average sapwood thickness of three inches would have about 84 per cent sapwood in the total volume. For the same concentration of preservative, it might be assumed that poles with a sapwood thickness of 0.75 inch would require only 33 per cent as much preservative as when the sapwood is three inches thick. It should be borne in mind, however, that there is a greater opportunity for more rapid loss of preservative from thin sapwood through leaching and evaporation because the preservative is concentrated closer to the surface. In the latter case, somewhat heavier absorptions would help compensate for this difference in depth of penetration.

Equation 2 will be found convenient for comparing the relative proportions of sapwood in a timber of any given diameter when different depths of sapwood are assumed.

Treating Conditions and Pressure Treatment Specifications

In addition to specifying the preservative, the average absorption and penetration required, and the method of seasoning or conditioning the poles for treatment, it is desirable to specify the maximum preservative temperatures and pressures to be used, since these have an important bearing on the success of treatment. Some species will withstand more severe treating conditions than others.

Conditioning

Green southern yellow pine poles commonly are conditioned for treatment by the steaming and vacuum process. Although the average amount of water removed is usually not more than about five to six pounds per cubic foot, the removal of this amount of water and the heating of the wood to a favorable treating temperature as a result of the steaming make this method a very satisfactory means of conditioning southern yellow pine poles. This method, nevertheless, is not well adapted for conditioning other woods commonly treated because either the other woods are injured more easily by the temperatures and heating periods required in steam conditioning or the final moisture content still may be too high for good penetration after the steaming treatment is applied.

Green coast Douglas fir poles are conditioned by the Boulton, or boiling under vacuum, process. In this process, round timbers are usually boiled under vacuum with a specified maximum preservative temperature of 220 degrees Fahrenheit. Since water is evaporated during the boiling period, the wood temperature is usually considerably lower than the temperature of the heating medium. This process also is applied in the conditioning of other species in the green condition and has the advantage that it removes water from green material under mild temperature conditions.

Air seasoning is, of course, the most widely used method of conditioning wood preparatory to treatment.

Treating Temperatures

Both laboratory experiments and subsequent studies made under commercial treating conditions have shown that treating temperatures of 190 degrees to 200 degrees Fahrenheit are much more effective in obtaining good penetrations than lower temperatures. This applies when either preservative oils or water solutions are used. Temperatures a little higher than 200 degrees Fahrenheit often can be used to advantage for some species, such as the southern pines.

Preservative Pressures

Preservative pressures should be kept low enough to prevent objectionable checking and collapse. Some species, as for example southern yellow pine and some of the hardwoods, can withstand treating pressures considerably higher than many other woods. Whenever objectionable checking or collapse occurs as a result of the temperature and pressure conditions employed, the preservative temperature should be maintained in the range of 190 degrees to 200 degrees Fahrenheit, and the treating pressure should be lowered as may be needed to prevent unnecessary checking or collapse.

Specifications

The American Wood Preservers Association for many years has had specifications covering the pressure treatment of southern yellow pine and coast Douglas fir poles, and these specifications have been revised at various times. In recent years this association has prepared specifications for the pressure treatment of western red cedar, lodgepole pine, jack pine, and red pine poles.

Treatment of Species Listed in the Various Groups

The recommendations made in the subsequent discussion are based on experiments made by the Forest Products Laboratory in a study of the treatment of the species named.

Species in Group I. Atlantic white cedar has a sapwood depth about the same as that of western red cedar (about one half to one inch for most timbers) and can be treated under the same conditions as specified for western red cedar (American Wood Preservers Association).

The spruces are resistant to penetration in both the sapwood and heartwood, and it is also difficult to distinguish the sapwood from the heartwood. Sapwood of the freshly cut timbers usually has a high moisture content in the range of 140 to 165 per cent or over. In order to avoid objectionable checking and collapse in the spruce species, it is desirable to limit the preservative pressure to about 150 pounds per square inch when the Rueping process is used and to about 120 to 130 pounds per square inch when the Lowry, or full cell treatment, is applied.[8]

Species in Group II. The sapwood of ponderosa pine usually ranges from about two to three inches in thickness, and the sapwood moisture content of the freshly cut wood is relatively high, averaging about 150 per cent. This is one of the few softwood species that are comparatively easy to treat in both the sapwood and heartwood when air seasoned.

The sapwood of the white pines commonly ranges from about one to two inches and occasionally to three inches in thickness.

In general, the treating conditions specified for the southern pines will be suitable for the pines listed in group II.

Species in Group III. Jack pine has a sapwood that commonly ranges from about $1^1/_2$ to $2^1/_2$ inches in thickness, although poles grown in some regions may have sapwood considerably less than one inch in thickness. Like lodgepole pine, this species is injured more easily by high pressures than some of the other pine species, and the maximum treating pressure should not exceed 150 pounds per square inch and in most cases should be somewhat lower. The treatment of poles of this wood is covered in the American Wood Preservers Association specification for the pressure treatment of jack pine poles.

Red (Norway) pine has a fairly deep sapwood that ranges from about two to four inches in thickness. This species has

a high average moisture content of about 135 per cent in the sapwood when the timber is freshly cut. The treatment of poles of this species is covered in the American Wood Preservers Association specification for the pressure treatment of red pine poles.

The white firs, like the spruces, are very resistant to treatment in both the sapwood and heartwood, and the sapwood is not distinguishable from the heartwood. Treating conditions previously discussed for the spruce species also will apply for the true firs, such as white fir (*Abies concolor*).

Species in Group IV. The sapwood of Douglas fir poles grown in the region between the Pacific Coast and the Cascade Mountains normally has a range in thickness of about $^3/_4$ to $2^1/_4$ inches. In most cases it will average between 1.3 and 1.4 inches. The moisture content of the sapwood in freshly cut timbers usually will average around 115 per cent. The treatment of the coast-type Douglas fir poles is covered in the American Wood Preservers Association specification for the pressure treatment of coast Douglas fir poles.

The mountain-type Douglas fir has a thinner sapwood than the coast type and is usually about one inch or less in thickness. The so-called Inland Empire or intermediate type that grows between the coast and the Rocky Mountain region has a sapwood thickness ranging from about one to two inches.

While the heartwood of the timber that grows in the Rocky Mountain region is far more resistant to treatment than the heartwood of the timber grown in the coast region, the sapwood of the mountain type can be treated satisfactorily by the pressure process. Although the mountain type wood has a lower mechanical strength than the coast wood, the timber grown in the mountain region will withstand a somewhat higher treating pressure without showing a marked increase in checking and collapse. Treating pressures as high as 140 to 150 pounds per square inch generally can be used for the Rocky Mountain wood, while it is usually desirable to use pressures 25 to 30 pounds lower for the coast material.

The sapwood of eastern hemlock is difficult to distinguish from the heartwood and, like the heartwood, is relatively resistant to treatment. The average moisture content of the sapwood of the eastern species is in the neighborhood of 120 per cent in freshly cut wood. Poles of this species should be air seasoned for treatment.

The sapwood of western hemlock is sometimes lighter in color than the heartwood and is usually not much over an inch in thickness. The average moisture content of the sapwood of the western species is considerably higher than that of the eastern species and averages around 170 per cent. Both hemlock species can be treated using pressures up to about 150 to 160 pounds per square inch.

The sapwood of both eastern and western larch is usually less than one inch in thickness; hence the sapwood thickness of these woods is similar to that of the cedars, Rocky Mountain Douglas fir, and western hemlock. The average moisture content of the sapwood of freshly cut western larch is in the neighborhood of 125 per cent.

Treating pressures as high as 175 pounds have been used in treating larch timbers. This species usually can withstand somewhat higher treating pressures than most of the softwoods grown in the western states.

The purpose of the foregoing outline discussing treating conditions and specifications is to point out some of the more important variables that should be considered in the preparation of specifications for pressure treatment. It should be borne in mind that if the best results are to be obtained, the treating conditions should be adjusted to meet the structural variability of the different woods.

Crossarms

The suitability of a given species for crossarms will depend largely on the strength properties and freedom from such defects as cross grain and knots.

Up to the present time, southern yellow pine and coast Douglas fir have been considered the standard woods for crossarm material. The southern pines usually have a larger proportion of sapwood, and since the crossarms of this species are used in large quantities in the South where service conditions are more severe, it is customary to pressure-treat them. The absorptions commonly specified range from around 8 to 12 pounds of creosote per cubic foot.

Coast Douglas fir crossarms generally are used without treatment because they normally have little sapwood and are used more widely in the northern and western states where decay conditions are less severe than in the southern states.

Conditions under which crossarms are used are generally less favorable to decay than where wood is in contact with the soil. For this reason the heavier absorptions used for timbers exposed to the more severe service conditions are not so important for crossarms. The shortage of southern yellow pine and Douglas fir arms that, as in the case of poles, has developed as a result of the war, has made it necessary to consider substitute species.

In the pole species listed under groups I to IV, the conifers that seem most promising for crossarm material include western larch, mountain or intermediate-type Douglas fir, western hemlock, and red pine.

Treatment of Crossarms

Species that have a considerable proportion of sapwood or that do not have good durability in the heartwood should be treated if good service life is to be obtained. Woods of this kind include western hemlock and the various pine species.

Crossarms made from western larch or the Rocky Mountain and intermediate-type Douglas fir would have little sapwood because of the thin sapwood in trees of these species. The heartwood of these woods is similar to the coast-type Douglas fir heartwood from the standpoint of natural durability. When heartwood crossarms are pressure-treated, it is probable that it would be difficult to keep the net retentions much below five to six pounds per cubic foot because of the large amount of surface area in proportion to the volume. Pressure treatment, however, should give deeper penetrations in heartwood material, and when there is a considerable amount of sapwood it should require somewhat lower net retentions than when the open-tank treatment is used.

One of the larger pole using companies has now revised its crossarm specifications to include open-tank-treated red pine, jack pine, lodgepole pine, and the inland or intermediate Douglas fir. Jack pine and lodgepole pine are somewhat lower in strength properties than the other species named, but it is assumed that they will provide suitable crossarms for the lighter service conditions. This company has treated several thousand arms of these woods using a hot bath of coal tar creosote and a cold bath of five per cent pentachlorophenol in an aromatic petroleum oil. These treated arms have been installed in service, and an examination of their condition is to be made at various periods.

Although the use of the pentachlorophenol solutions is still in the experimental stage, it would seem that an open-tank treatment, such as that mentioned, should give good results. The open-tank treatment of crossarms appears to be the most suitable substitute treatment where pres-

Electric Equipment for Cornell Variable Density Wind Tunnel

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Synopsis: Spectacular advances in airplane design during the past decade have created an increasing demand for additional testing facilities with which to obtain those data essential to successful design. To answer partially this demand and to facilitate more basic aerodynamic research on the subject of compressibility effects experienced in high speed flight, several new wind tunnels were constructed during this same period. These tunnels vary widely in size and in requirements for air density and air speed in the tunnel working section. Correspondingly, the horsepower output requirements for their main fan drive motors vary widely. But in all cases close speed regulation at preselected speed values over a wide range is essential to good test results. Most modern tunnels also incorporate suitable electric equipment for the operation of powered models and for the calibration of motors used in these models. This paper describes the Cornell variable density wind tunnel from the viewpoint of electrical engineers and discusses principles governing the selection of electric equipment for wind tunnels.

THE WIND TUNNEL with details as shown by Figure 1 is the variable density high speed tunnel recently constructed by the Curtiss-Wright Corporation Research Laboratory (now Cornell Aeronautical Laboratory), Buffalo, N. Y. This project was initiated by Doctor Norton B. Moore, then of the Curtiss-Wright Corporation, and construction was started under his direction in 1942. The basic design of the tunnel was prepared by members of the staff of the California Institute of Technology under the supervision of Doctor Clark B. Millikan. This group also acted as consultants during construction of the new facility. The design was financed jointly by the Curtiss-Wright Corporation and by a group of four Southern California aircraft companies which finances and owns a basically similar facility, known as the cooperative wind tunnel, in Pasadena, Calif.

The tunnel consists of a reinforced steel duct of circular cross section, housed in a brick-faced frame building, which adjoins a laboratory and power plant building. The principal dimensions of the duct, which is built of steel plate approximately seven-eighth inch thickness, are shown in Figure 2. The duct forms a closed circuit through which air is circulated by the fan system indicated as *A* in Figure 1. With the outlet door of the tunnel closed the static pressure of the air in the tunnel may be varied between one-fourth atmosphere and four atmospheres, and then the air may be caused to circulate in this condition. Vanes built into each corner to guide the air flow are shown in Figure 2. Cooling of the air is accomplished by a bank of finned coil radiators mounted on the downstream side of the turning vanes in one corner of the tunnel. This is shown at *B* in Figure 1. Cooling water is pumped through these radiator coils and the heat is transferred to the atmosphere by means of the cooling tower shown at *C* in Figure 1. This cooling system is of adequate capacity to remove heat corresponding to the maximum continuous power input to the fan system, while limiting the air temperature inside the tunnel to 125 degrees Fahrenheit.

The airplane model shown as *D* in Figure 1 is supported in the "test section" of the tunnel within a sphere, which can be isolated from the rest of the tunnel duct by means of lock gates. This arrangement, unique to this tunnel and the similar one at the California Institute of Technology, Pasadena, allows pressure to be maintained in 90 per cent of the total tunnel volume, while the sphere is isolated decompressed, and the model serviced or changed. Considerable effort was made in the design to speed up this process, thus reducing to a minimum the time during which the tunnel is out of service.

In Figure 2, in the plan view of which the air flow is counterclockwise, the contraction leading into the "working section" and the diffuser downstream of the working section are shown. The inside diameter of the duct at the beginning of the contraction is 31 feet 6 inches, the cross-sectional dimensions of the working or experimental area are 12 feet wide by $8^1/_2$ feet high, and finally the inside diameter of the duct at the end of the diffusion area (ahead of the first corner downstream of the working section) is 17 feet 9 inches. The airplane model is mounted on supports which, when the tunnel is operating, are integral with a basic metrical (force measuring) system, to which the forces and moments produced on the model by the air stream thus are transmitted. These forces and moments are transmitted hydraulically to a "control" room for indication and recording. For servicing or changing of a model, a transfer cart, the top of which forms the floor of the working section, is raised and carries with it the model supporting struts. Raising of the transfer cart floor disengages the support system from the metrical system and provides clearance for the transfer cart to be motored out of the tunnel on steel transfer rails. Interchangeable transfer carts make it possible to set up completely and "instrument" one model in the model preparation room while a second model is in the tunnel under test.

The fan system, mentioned previously, is driven through a 30-foot hollow steel shaft by a 10,250-horsepower continuous rated 2-unit motor set. Speed adjustment of this drive over a range from 0 to 570 rpm, with close regulation at preset values between 50 and 570 rpm, is arranged for remote operation from the control room near the tunnel door. Power supply for the complete installation is taken from the 115,000-volt 3-phase 60-cycle lines and stepped down to 4,800 volts before it enters the power area building and is connected to the metal-clad switchgear. Figure 13 shows the operators' console for the Cornell variable density wind tunnel.

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sure treatment with creosote or with solutions of creosote and pentachlorophenol is not feasible or does not seem necessary for the conditions of service to which the crossarms are exposed.

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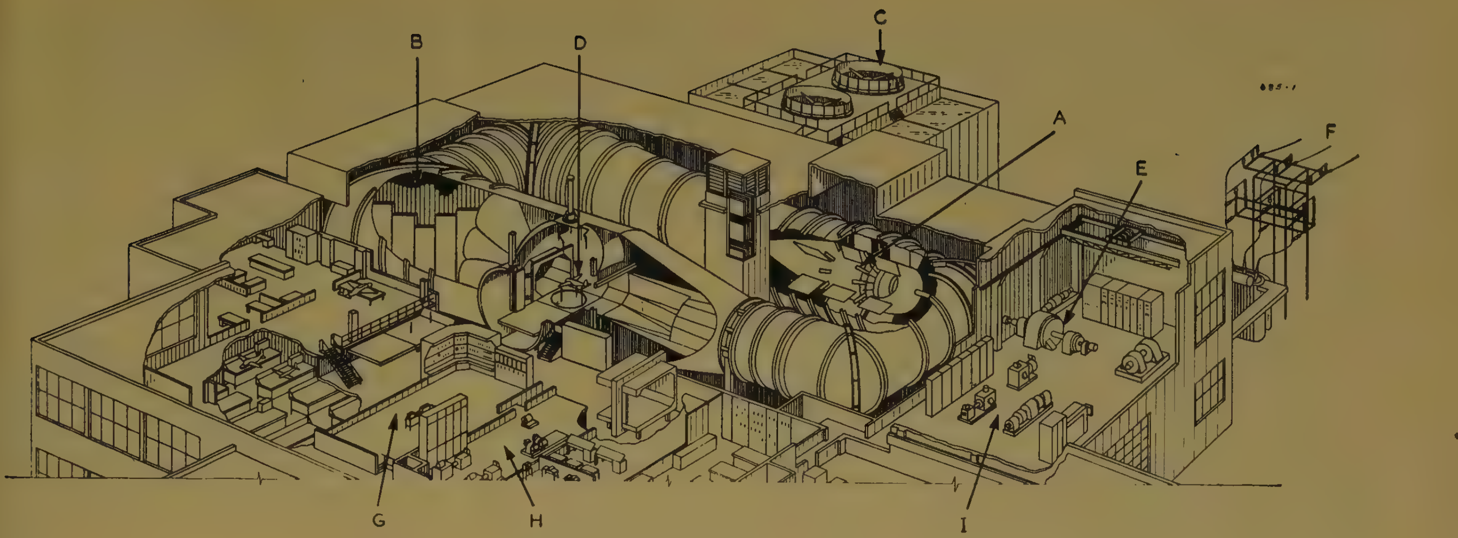


Figure 1. Cutaway view of Cornell Aeronautical Laboratory showing wind tunnel installation

A—Fan impeller
B—Heat exchanger
C—Cooling tower
D—Airplane model in test section
E—10,250-horsepower driving motor
F—115,000-volt 60-cycle power supply
G—Control room
H—Dynamometer room
I—Adjustable frequency system

For operation of powered models in the working section of the tunnel, an adjustable frequency system with frequency adjustment over a range from 60 to 450 cycles per second is arranged for remote operation from the control room. Precision metering of this variable frequency power supply is provided for indication and recording in the control room. The motors used in the powered models are calibrated on induction-type dynamometers capable of testing model motors up to speeds of 27,500 rpm, and up to maximum horsepower ratings of 200. Additional details of the electric equipment briefly mentioned in the foregoing will be given later in this paper.

Aerodynamic Principles Determining Power Requirements of Tunnel

Some knowledge of the aerodynamic principles influencing the design details of the tunnel is necessary for correct application of electric equipment on an installation of this type, although it is the intention of this paper to dwell on the electrical phase of the subject.

The primary object of all wind tunnels is to perform tests on scale models, which tests will produce data to assist the airplane designer in his design of new airplanes or in modifications of existing designs. For this purpose the wind tunnel is designed to measure accurately the reactions on the model at various altitudes and in an adjustable velocity uniform air flow. Various forms of tunnels have been designed[1] to produce effectively this adjustable uniform air flow through an experimental section in which the model is mounted, but the majority of recently designed tunnels built for testing in air velocities up to the speed of sound follow the general geometry of that shown in Figure 2. Force and moment reactions on the airplane model (and the coefficients derived from these) correspond to similar information on the full-size airplane, only if the model is geometrically similar to the full-scale design and also only if the model test is made at a Reynolds number equal to that expected on the full-size airplane in flight. Remembering that Reynolds number

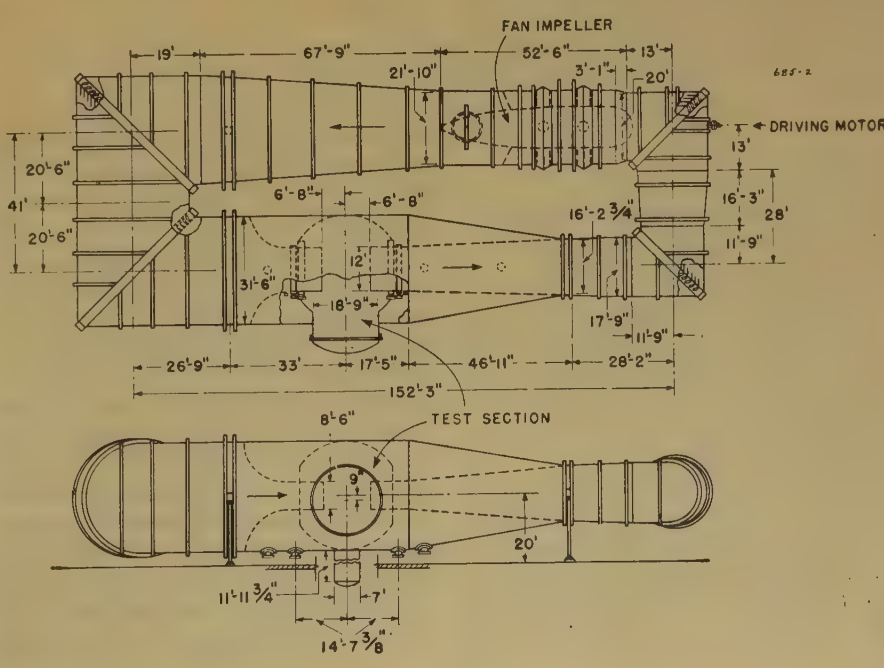
$$R=\frac{Vl\rho}{\mu} \qquad (1)$$

in which

V = relative velocity between body and fluid
l = a characteristic dimension of the body
ρ = mass per unit volume of the fluid
μ = coefficient of viscosity of the fluid

it can be seen that for testing in an atmospheric pressure tunnel, the Reynolds number for flight conditions can be equalled only by using a full-scale model, or alternatively testing a fractional scale model at higher velocities than flight

Figure 2. Outline of wind tunnel structure



conditions. In the case of airplane testing, velocities above that of flight conditions may introduce other more serious errors because of the compressibility of the air and, consequently, the testing of a full-size model is dictated. Some wind tunnels for full-scale testing are in operation, but their initial costs and the power required for their operation are so high that, to date, they have been installed only by centralized government agencies. The majority of wind tunnel testing is done on fractional scale models at reduced Reynolds number values; full-scale conditions are extrapolated from these data.

Inspection of equation 1 indicates that change of density causes a corresponding change in the Reynolds number. The density is proportional to the air pressure, and the viscosity is approximately independent of pressure, so that Reynolds number may be changed approximately proportional to the pressure. A wind tunnel taking advantage of this feature was first built and operated by the National Advisory Committee for Aeronautics at Langley Field, when a relatively small tunnel operating with compressed air at 20 atmospheres pressure was put into operation in 1927. This feature also was incorporated in the design of the tunnels recently built by the Curtiss-Wright Corporation (now the Cornell Aeronautical Laboratory), the California Institute of Technology, and by the National Advisory Committee for Aeronautics at Ames Aeronautical Laboratory, Moffett Field, Calif.

Consider now the power requirements for wind tunnels: Corresponding to a specific air velocity through the working section (throat) of the tunnel, the kinetic energy of the air stream passing through the throat is given by the equation:

$$KE = 1/2\rho_o V_o^3 A_o \qquad (2)$$

Where

ρ_o = mass density of air in the throat
V_o = rated air velocity at throat
A_o = cross-sectional area of throat

Assuming approximately uniform velocity distribution throughout, the summation of energy loss in the various portions of the tunnel duct may be expressed in the form

$$\text{energy losses} = K_o 1/2\rho_o\, A_o\, V_o^3 \qquad (3)$$

where K_o is the basic loss coefficient. With rated velocity conditions, this is the energy which must be supplied by the fan and, taking into account fan efficiency ηf, the fan shaft

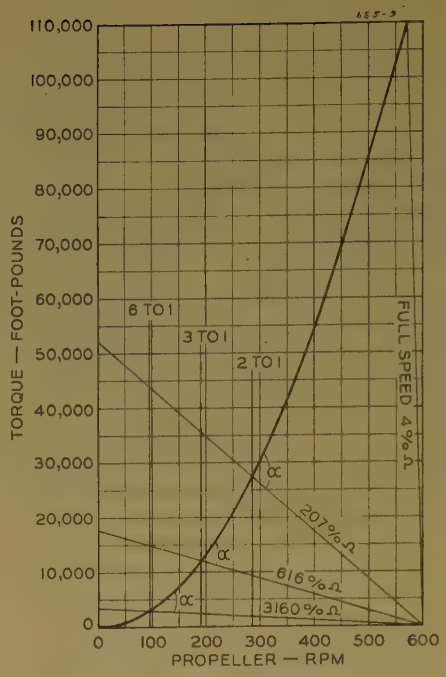


Figure 3. Speed-torque characteristics of 12,000-horsepower 570-rpm fan load

Values of secondary resistance required with a single speed slip ring motor for 2:1, 3:1, and 6:1 speed reduction also are shown

power input is given by the equation

$$hp = \frac{K_o 1/2\rho_o A_o V_o^3}{\eta f \times 550} \qquad (4)$$

From equation 4 it can be seen that for a fixed throat cross section, the horsepower delivered from the main fan drive will be

$$hp = \frac{K\rho_o V^3}{\eta f} \qquad (5)$$

in which

$$K = \frac{K_o \times A_o}{2 \times 550}$$

and from which will be found an explanation for reducing tunnel static pressure (and therefore ρ_o) sufficiently to increase materially the air velocity in the throat without increasing the power input. For example, assuming that fan efficiency is maintained (by pitch change) a reduction from one atmosphere to one-fourth atmosphere in static pressure allows a velocity increase of approximately 58 per cent without increasing the horsepower input to the fan. This ability to study high velocity performance on models, without increasing fan input horsepower, prompted designers to provide for static pressure reduction in several modern wind tunnels. This equation also indicates that an increase in air pressure (and corresponding increase of Reynolds number) calls for a corresponding increase of fan shaft horsepower or, alternatively, air velocity at the throat will be reduced if horsepower input to the fan is held constant. For example, by equation 4, if the density is increased from atmospheric to that at four atmospheres pressure, the velocity at the throat will be reduced to approximately 63 percent of that at atmospheric pressure, if horsepower input is held constant.

Summarizing, the change of static pressure above and below atmospheric pressure has provided a means of increasing the Reynolds number and the air velocity, respectively, without requiring a proportionate increase of model size or fan horsepower input.

Selection of Main Drive

Wind tunnel power plants can and do take many forms, depending upon the type of tunnel to be powered, and the purchaser's preference. In this particular instance we are considering electric drive, and as the tunnel is of the two dimension or variable density type, with single rotation, the propeller drive unit will be mounted outside the tunnel proper—a construction detail which simplifies greatly the problem of the motor design. Also, power requirements are quite moderate, therefore permitting the use of any of the various types of drives.

The slip ring induction motor is perhaps the least expensive variable speed electric motor unit. This machine, with resistance in the secondary winding, has series characteristics, the speed being a function of secondary resistance and shaft torque.

The conventional propeller system has a torque characteristic which varies as the square of the speed for any established blade angle and air density. At rated speed, torque is, of course, 100 per cent, while at half speed it is 25 per cent. Usually the minimum requirement for speed range is 6 to 1, with the torque requirement at one-sixth speed equal to approximately three per cent. This is, of course, far outside the practicable operating range of a single speed slip ring motor. As a matter of fact, it taxes the performance of a two speed or pole changing motor with a two to one synchronous speed, and a horsepower range of the order of 8 to 1.

Such two speed drives have a point of discontinuity at the speed change point. They also require two slip regulators. With induction motors the speed varies

with change of applied voltage for a particular setting of secondary resistance and the propeller torque varies with changes of the model attitude, or change of model size in the test section. Corresponding to a change in load torque the induction motor speed will change, and the amount of change varies widely according to the amount of resistance in the secondary circuit. This type of motor inherently operates at low power factor, especially when lightly loaded. Because of these characteristics this combination of machines has not found complete acceptance on the part of wind tunnel operators. Economics, of course, play a major role in any large program of this nature, and the consultants for the subject project wished to retain the economies of the induction motor drive for maximum power requirements, and supplement the single speed induction motor with some different type of speed control between minimum speed and half speed.

One of the fundamental tunnel specifications was a 10 to 1 speed range which obviously could not be met through the use of the slip regulator, as will be evident from the speed torque requirements of a 12,000-horsepower 570-rpm propeller system (Figure 3). The parabola, of course, is the propeller torque curve, while the straight lines are the speed torque curves of the induction motor with various secondary resistances, with resistance values in per cent. The angle of the intersection (α) decreases rapidly below 50 per cent speed. The slip regulator, however, should function successfully between half speed and full speed.

Various systems were studied, but it was found to be most economical, for the rating involved, and yet fully reliable, mechanically, to combine the two basic adjustable speed drives, that is to combine an adjustable voltage (Ward Leonard) drive, with the slip ring motor and slip regulator to form a tandem drive. The adjustable voltage drive is operating alone over the range of 10 per cent to 45 per cent speed, while between 45 per cent and 100 per cent speed, both drives share the load: the d-c equipment providing the vernier control, so to speak, and the induction motor with its slip regulator the coarse adjustment.

The basic power requirement for the tunnel was set at 12,000 horsepower at 570 rpm for 30 minutes, the use of adjustable pitch propellers permitting this specification to fit into any tunnel pressure or air density. The power requirements of the conventional propeller vary as the cube of the speed. Thus, at half speed, or 285 rpm, the torque is reduced to 25 per cent rated, and the horsepower to 12.5 per cent or 1,500 horsepower.

The d-c part of the drive, as noted previously, provides the entire power to the propeller up to 45 per cent of rated speed, or 260 rpm. The motor rating of 1,250 horsepower (continuous) is more than ample to meet this requirement. Speeds

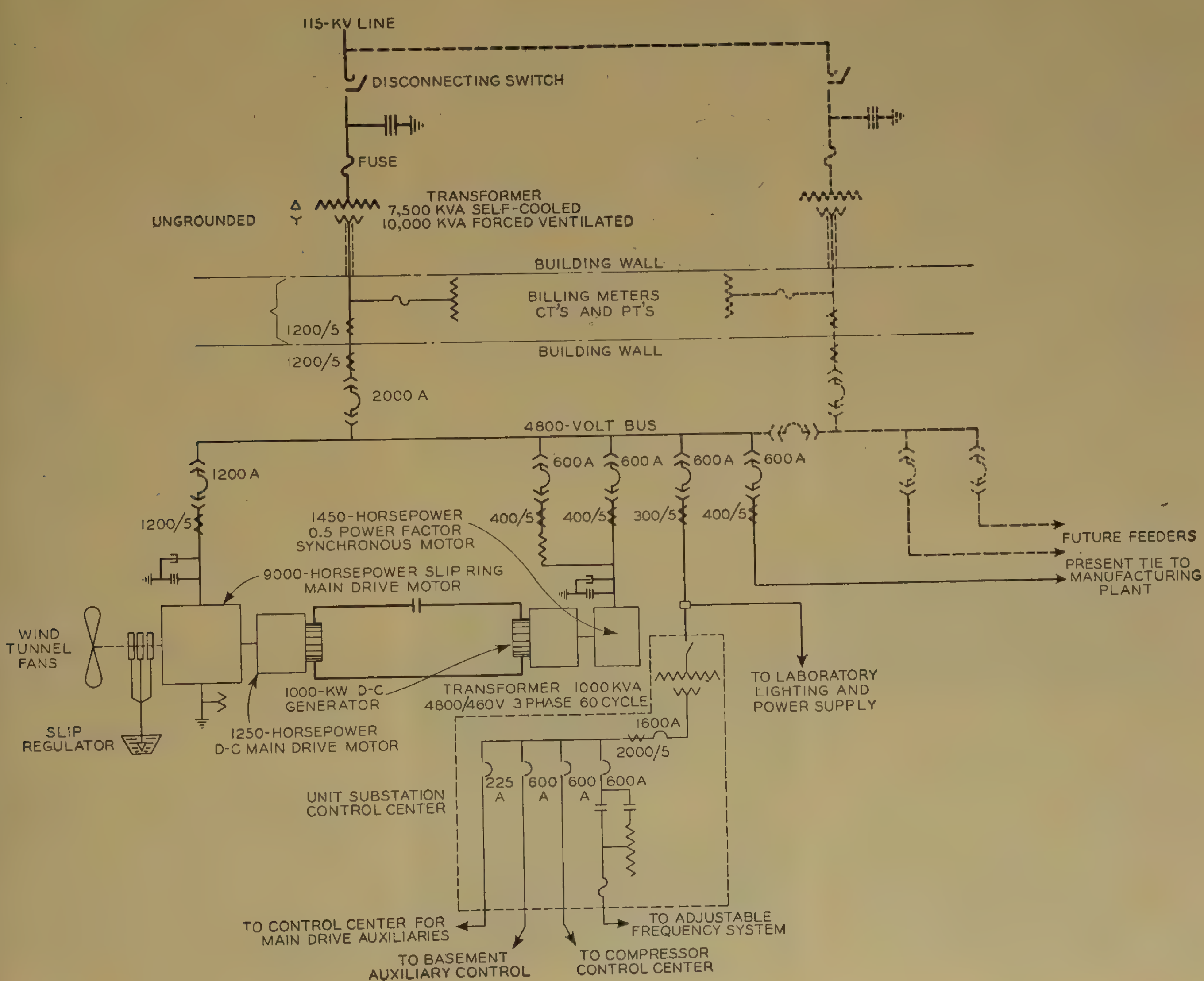
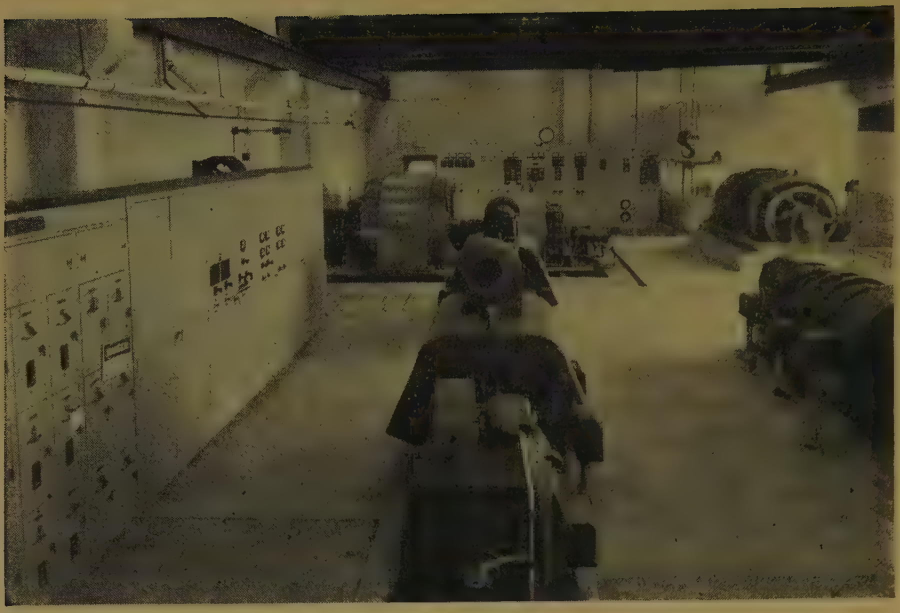


Figure 4. Schematic diagram of power circuits of the fan drive of the Cornell Aeronautical Laboratory



higher than 260 rpm are provided by motor field weakening, at constant generator voltage; that is, constant horsepower (1,250 horsepower) is available throughout the range 260–570 rpm.

Power to this d-c motor is derived from a 1,000-kw motor generator set operating from the 4,800-volt 3-phase 60-cycle power supply. Generator voltage (Ward Leonard) control is used to obtain motor speed range 0–260 rpm. Obviously, the available horsepower then is reduced in proportion to generator voltage. The d-c motor and the motor generator are designed for continuous duty at their nameplate rating and are capable of carrying short-time loads of 200 per cent rating (that is 2,500 horsepower at full voltage).

The induction motor was rated on a continuous basis of 9,000 horsepower at 590 rpm and is capable of delivering 11,250 horsepower for 30 minutes. The tandem unit has a total capacity of 12,500 horsepower for 30 minutes, with ability to handle peak loads corresponding to a torque of 15,000 horsepower (torque) at 570 rpm. Figure 4 shows the general schematic diagram for this drive, the arrangement of the machines, and the power supply. Figure 5 shows the tandem drive unit installed.

Figure 5. Motor room of Cornell variable density wind tunnel

In background is 4,800-volt metal-clad switchgear and main drive control. In center is 10,250-horsepower continuous rated tandem motor set and 1,000-kw motor-generator set. In foreground are adjustable frequency system machines and on left its control and control center for all motor room auxiliaries

Speed and Load Control

The d-c part of the drive serves as a means of controlling the speed and of maintaining it at the value preset by a master rheostat. The slip regulator is used for controlling the division of load between the d-c motor and the induction motor. A schematic diagram indicating the speed and load control system is shown in Figure 6.

The drive is started from rest by generator voltage control using the 1,250-horsepower d-c motor only; the induction motor remains disconnected from its power supply during this starting period. In this manner speeds up to 260 rpm are obtained. Above this speed the power requirements of the fan exceed the rating of the d-c machines, and the induction motor is connected automatically to the line with the slip regulator in its "all resistance in" position. The "load control" responsive to the direct current in the armature circuit of the 1,250-horsepower motor causes the slip regulator to change the secondary resistance of the induction motor until the latter assumes load such that the a-c motor and the d-c motor each are loaded to approximately the same percentage of their rated capacity.

The load proportioning is effected by use of an amplidyne regulator. Stated briefly, it operates in the following manner: A voltage proportional to the d-c motor armature current is compared with a voltage which varies inversely as the height of the slip regulator electrodes. The difference of these two voltage signals is applied to the field of an amplidyne generator. The armature of this amplidyne is connected to the armature of a separately excited shunt wound d-c motor driving the slip regulator electrodes up or down. This increases or decreases, respectively, the secondary resistance of the induction motor causing it to transfer load to or from the d-c motor until the two signal voltages are equalized, and the electrode motor stops.

The speed of the drive is maintained by means of an electronic-amplidyne regulator as follows: The drive is equipped with a permanent (Alnico) field tachometer generator. Its voltage is compared with a potentiometer connected across a regulated 125-volt d-c reference source. The position of this potentiometer determines the speed at which the drive operates.

The difference between the voltages of the tachometer generator, on the one hand, and the potentiometer, on the other hand, is applied to the grid of an electronic tube. Thus amplified, this voltage then is applied to the regulating field of an amplidyne exciter. The latter controls the voltage of the 1,000-kw d-c generator, and regulates speed to preset values throughout the speed range. Thus, if the tachometer voltage is lower than it should be, the drive is speeded up until the proper speed level is reached.

An amplidyne exciter also supplies the field of the 1,250-horsepower d-c motor. The field current of this motor is held constant for speeds between 0 and 260 rpm and is preset at lower values (by the master rheostat) for speeds between 260 and 570 rpm.

Power Factor Control

As will be appreciated readily, a large induction motor operating at light load would have a very poor power factor, drawing in this instance about 2,200 reactive kva. To bring this power factor up to 95 per cent minimum (lagging) for all conditions of operation, the synchronous motor on the motor generator set has a rated output at 50 per cent power factor, leading, and a power factor regulator was added to the control. Thus, when the induction motor is connected, the power factor regulator automatically comes into play, the leading kilovolt-amperes of the synchronous set counteracting the lagging kilovolt-amperes of the induction machine.

This regulator is responsive to the power factor in the incoming 4,800-volt feeder supplying both the 9,000-horsepower motor and the 1,000-kw set and controls the excitation of the 1,450-horsepower synchronous motor to maintain this power factor at the high value mentioned. Figure 6 shows schematically the

elements of this system and indicates that the 1,450-horsepower synchronous motor field actually was supplied by a conventional d-c exciter. This in turn is excited separately by an amplidyne exciter, the field of which is controlled by the power factor regulator.

Main Power Supply

The availability of an adequate power supply is an important consideration in determining the location of an installation of this type.

The power system must be selected to handle high intermittent loads corresponding to full power running of the tunnel, although total energy consumption is comparatively small. For example, the Cornell tunnel will have 20-minute peaks of approximately 10,000 kva which will occur when running the tunnel at top air speed. Such tests would consume perhaps ten per cent of the running time of the tunnel, while the majority of tunnel tests will have power demands of less than half this value. It will be remembered that the power required varies as the cube of the air speed; for example, a half speed test in the tunnel requires only one eighth the power of a full speed run.

An adequate power supply for the Cornell tunnel is provided by a short overhead stub feeder from the 115-kv 3-phase 60-cycle transmission line of the Niagara Lockport and Ontario Power Company. The laboratory end of the feeder line is terminated on an outdoor substation structure. This structure mounts 115-kv manually-operated disconnect switches, thyrite lightning arresters, and boric acid fuses, connected ahead of the transformer as shown by the schematic diagram in Figure 4. The transformer is a 3-phase oil-immersed unit rated 7,500-kva self-cooled, or 10,000-kva forced ventilated. It steps down the voltage from 115 kv to 4,800 volts. The ventilating fans are started automatically as a function of transformer temperature or whenever the tunnel fan drive is run above a predetermined speed.

The secondary winding of the transformer is throat connected through the wall of the wind tunnel motor room, at first floor ceiling height. After passing through a secondary metering room the 4,800-volt copper tube bus connection is carried up through the first floor ceiling directly into an incoming feeder unit of a metal-clad switchgear line-up. Figure 4 shows, diagramatically, the electrical layout of the 4,800-volt switchgear, the air circuit breaker ratings, and the provision for future extension of the system.

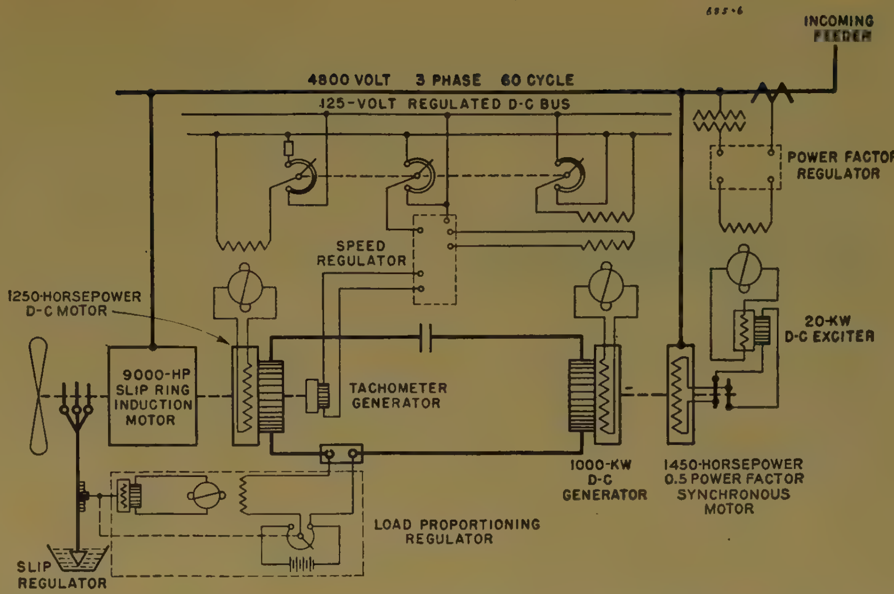


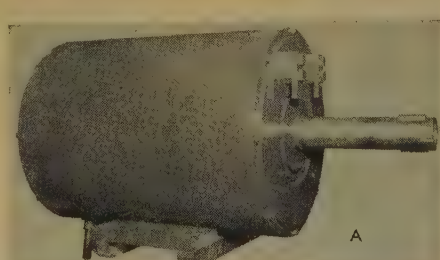
Figure 6. Elementary diagram showing speed-load, and power factor regulating features, incorporated in control of fan drive

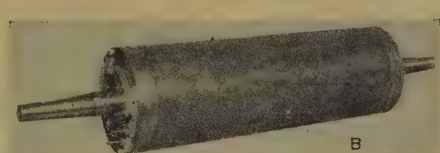
Powered Model Testing

It has been pointed out how the force acting on a model airplane may be measured in the test section of the wind tunnel. The lift and drag, of course, are determined readily, but the behavior of the airplane under the resultant forces imposed by its power plant have not been determined. These forces may be provided by building a power plant in the model airplane. An electric motor is used in place of the gasoline engine. The model, of course, will be an exact scale model. If this model is one tenth the size of full-scale model, the propeller will be one-tenth the diameter of the actual propeller. Its speed, therefore, must be ten times the speed in order to maintain the model propeller tip speed at the same value as that of the full-scale propeller. If the rated speed of the actual propeller is 900 rpm, the model must run at 9,000 rpm, and power requirements are reduced inversely as the square of the scale.

The high frequency motors used for this purpose are quite special. Space limitations are usually quite severe. Consider, for instance, the one-tenth scale model. If the nacelle frontal diameter is 40 inches, the model can be but four inches. If the output of the full-scale engine is 1,800 horsepower, the model must be 18 horsepower. It is then no mean task to build an 18-horsepower motor with a diameter of four inches. Motors of this type are water-cooled; that is, the frame includes a water jacket. Air cooling is not practicable because the discharge of air produces a jet effect, and this would affect the accuracy of the results.

Model motors are almost without exception tailor-made. Speeds as high as 80,000 rpm at 0.1-horsepower $1^3/_4$-inch diameter have been provided, and horsepowers as high as 3,000 at 2,400 rpm at the other extreme with a diameter of 29 inches. Figure 7 shows some of these machines. A motor of 300 horsepower at 20,000 rpm also has been built, in the type of construction shown. Figure 8





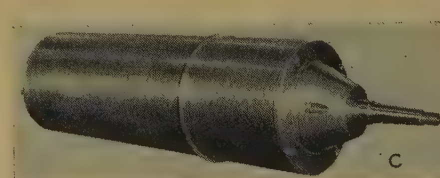


Figure 7. Water-cooled induction motors for use in airplane powered models

A. 1,000 horsepower, 2,200 rpm, 28 inches in diameter by 39 inches long

B. 200 horsepower, 5,000 rpm, 10 inches in diameter by 21 inches long

C. 250 horsepower, 10,800 rpm, 9 inches in diameter by 19 inches long

shows a counterrotation unit. This unit consists of two motors built in tandem; one has a hollow shaft. This is provided for counterrotation propeller studies.

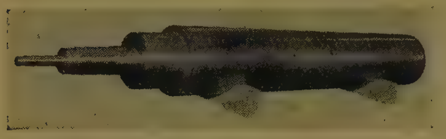


Figure 8. Contraturning water-cooled induction motor for airplane powered model

Hollow shaft unit (front) and solid shaft unit (rear) each are rated 75 horsepower at 3,000 rpm. Over-all diameter of motor is ten inches

Such speeds require a power supply at high frequency. These frequencies may be obtained with alternators, or frequency converters. Reference to Figure 9 will show, at the top, a 5-unit set driven by a 400-horsepower 1,200-rpm synchronous motor, operating from the 440-volt 60-cycle supply. The alternators *A* and *B*, each rated 156 kva, 440 volt, 60 cycle, furnish power to frequency converters *A* and *B*, respectively. The d-c generators *A* and *B*, each rated 100 kw, 250 volt, supply power to 140 horsepower 960/-1950-rpm d-c motors driving, respectively, these converters *A* and *B*. By varying the voltage of generators *A* and *B*, the speed of the converters *A* and *B* is adjusted. In this manner, the converters, receiving electric power at 60 cycles at their collector rings, change it to 60/450 cycles at their stator output. This output is used to operate the model motors at the required high speed. The voltage of this output is regulated by controlling the field of the corresponding alternator *A* or *B*.

With two independent frequency converters available, it is possible to operate them either in parallel, or separately, at different frequencies. This arrangement also readily permits studies in counterrotation of model motors and propellers. Counterrotation, as is well known, permits the use of smaller propellers, that is propellers with less sweep, and the torque on the airplane balances out. Although, in practice, contrarotating propellers have been driven, through gears, by a single power plant, the two independently adjustable power supplies described in the foregoing permit the study of counterrotation at various ratios between the two propeller speeds. The two independent power supplies also make it possible to perform a test in the tunnel at the same time as calibrating a second model motor in the dynamometer room.

The capacity of the adjustable frequency system is required to be adequate for operation of all powered models which may be tested in the tunnel. In selecting the specified capacity for each independent half of this system, the tunnel designers took into consideration the type and scale of airplane models to be tested. As will be appreciated from earlier discussion of powered models, the horsepower capacity of the model motor goes up but the frequency of its supply must come down, as the scale of the model is reduced. Also affecting the capacity of the adjustable frequency system required at various frequencies is the fact that both single propeller and multipropeller models will be tested. To satisfy these requirements the designers specified that each frequency changer should deliver 125 kw at 150 cycles and 50 kw at 450 cycles. The induction frequency converter designed to meet the first requirement has a capacity of 375 kw at 450 cycles. While each converter is inherently oversize at the high frequency, the driving motor of each frequency changer set can be selected to meet more exactly the requirements without excess capacity. The d-c driving motor of each frequency changer set is rated 140/75 horsepower 950/1,950 rpm at 250 volts. It is started from rest by generator voltage control, and operated in this manner up to 950 rpm. The frequency output of the converter which it drives varies correspondingly from 60 cycles to 250 cycles. Above this point the generator voltage is maintained at 250 volts, while the d-c motor speed is increased to 1,950 rpm by weakening its field. Frequency of the converter output varies correspondingly from 250 to 450 cycles. Under these conditions of operation the d-c motor load increases from 0 to 140 horsepower at 950 rpm, then decreases gradually from this speed to 75 horsepower at 1,950 rpm.

As pointed out in the foregoing, large model motors are required to operate at low frequency, and in order to keep the motor leads as small (and therefore as flexible) as possible it is normal to design the windings of these motors for a higher voltage per cycle than the higher speed smaller motors. On the other end of the range extremely high-speed small motors operating at, perhaps, 450 cycles, require a winding designed for low voltage per cycle in order to avoid excessively high voltages and mechanically fragile wire. Although national standards are not established, normal practice is to design model motor windings for operation at one of the following values of volts per cycle: 0.625, 1.25, or 3.75. This accepted practice imposes a further requirement, of wide voltage range, on the adjustable frequency power supply. To satisfy this requirement each frequency converter is arranged for reconnecting the stator winding to operate either as a single circuit delta or double circuit delta machine.

This power supply equipment is unique in that electronic control is used on the generator field, motor field, and the converter exciter (alternator) field. Such control is new, of course, only in this particular field, it having been applied successfully on wind tunnel main drives where very special characteristics were desired. It also has been used on many other industrial drives. This type of control saves space and lends itself readily to the extreme flexibility needed for a laboratory tool of this type.

A unique feature incorporated in this installation, and made possible by the electronic control, is a volts-per-cycle regulator of the converter output. This is of extreme value in the operation of model motors because of the need to operate nearer to the flux saturation point in this design of motor

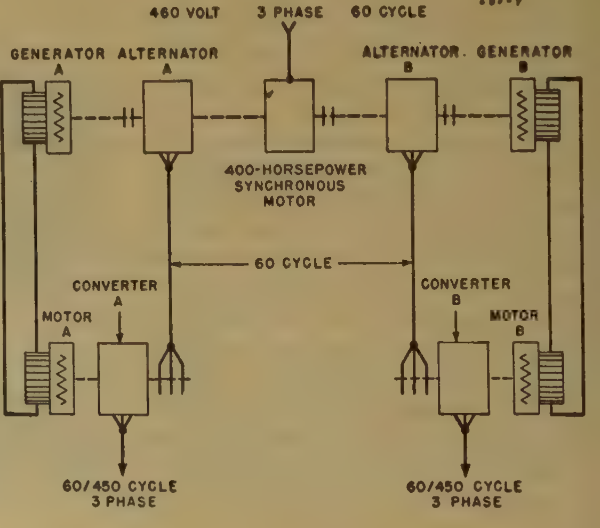


Figure 9. Schematic diagram of adjustable frequency supply system

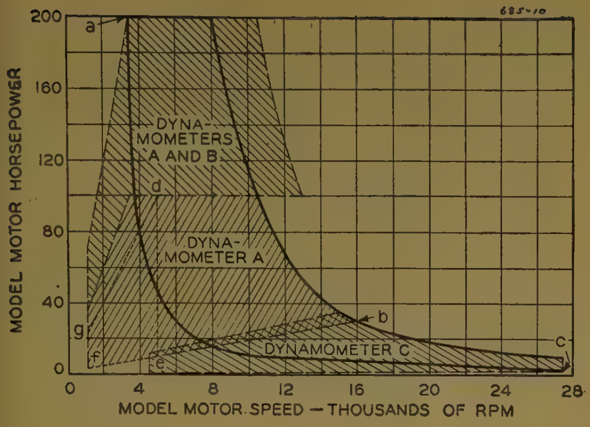


Figure 10. Horsepower-speed envelope of model motor maximum ratings to be tested on dynamometers, and coverage provided by dynamometers

than in more conventionally designed induction motors.

Model Motor Calibration

When testing a powered model in a wind tunnel it is, for aerodynamic reasons, necessary to control accurately the power delivered to the model propellers, according to the lifting force on the airplane model, as indicated by the wind tunnel balance system. The power input to the propellers may be determined either by direct torque and speed readings or by deduction from the electric power input to the model motors. The Cornell installation is equipped to use the latter method. In this method it is first necessary to calibrate the model motors for mechanical power output versus electric power input over the full range of torque and speed through which the motors can operate safely. A cradled dynamometer to perform this calibration for a specific motor presents no new problems but the selection, design, and manufacture of dynamometer equipment which will calibrate accurately all model motors to be used in one wind tunnel calls for many special features.

The selection of adequate capacity, range, and flexibility in the dynamometer installation must depend largely on the aerodynamicists' advice of what powered models can be tested in the tunnel, and his predictions of what will be tested. Based on such information, the design consultants for the Cornell tunnel defined an envelope of maximum horsepower and speed ratings of model motors to be calibrated on the dynamometer installation. This envelope is shown in solid lines in Figure 8. Specifications for calibrating equipment also required that, for any specific motor with maximum rating falling anywhere within the afore-mentioned envelope, the motor maximum torque be indicated accurately to within plus or minus one-quarter per cent. For a model propeller driving motor the maximum torque and maximum horsepower output will occur, of course, at maximum speed. For torque measurement below the maximum output it is desirable to indicate torque with an absolute error not greater than that corresponding to one-quarter per cent of the torque at maximum output of the motor on test.

It is also necessary to be able to calibrate all motors over a wide range of speed and torque. In Figure 10, a point *d* represents the maximum output of a motor for which calibration is required. The area *d, e, f, g*, indicates the full range of speeds and torques for which calibration of this motor is required.

Consideration of a few points in the envelope shown in Figure 10 gives a better conception of the total range of torques for which calibration equipment must indicate accurately. Consider the 200-horsepower 3,500-rpm model motor represented by point *a*. Its torque at maximum output is approximately 300 foot-pounds and it should be calibrated down to 30 foot-pounds torque. Then the 35-horsepower 16,000-rpm model motor is represented by point *b*. Its torque at maximum output is approximately 11.5 foot-pounds and it should be calibrated down to 1.15 foot-pounds torque. Finally the 2-horsepower 27,500-rpm motor is represented by point *c*. Its torque at this maximum rating is approximately 0.38 foot-pounds and it should be calibrated down to 0.038 foot-pound torque.

It is desirable to calibrate completely a model motor, throughout its speed and torque range, on a single dynamometer or, at most, on two dynamometers. On the other hand, the requirement of high accuracy over such a wide range of torques and speeds makes it desirable, from a design standpoint, to use several dynamometers to cover the range. In selection of machines to cover these requirements, consideration was given first to eddy current inductor-type dynamometers. In this design, mechanical power input to the dynamometer is converted into eddy current losses within the machine. These losses are removed from the machine by cooling water conducted to and from the cradled machine by flexible hoses. It is obvious that this type of machine can absorb only mechanical power and that the minimum torque which it can measure is that required to overcome friction and windage losses within the dynamometer. Although machines of this type are suitable for model motor testing from the standpoint of speed and required capacity, friction and windage are higher than desirable from the viewpoint of precision testing. Also, the necessity of cooling water hose connections is a handicap to attaining the required accuracy of torque measurement on this installation.

On the basis of using three dynamometers to cover the complete envelope of

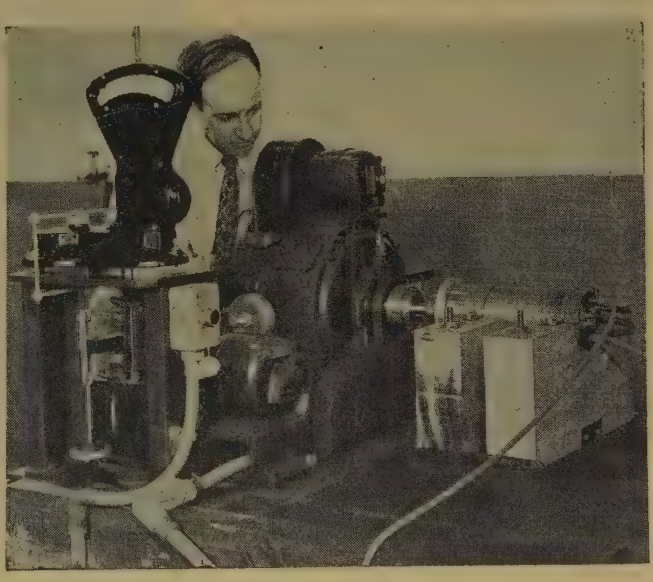


Figure 11. Water-cooled model motor being calibrated on a 35-horsepower 15,000/27,500-rpm induction-type dynamometer

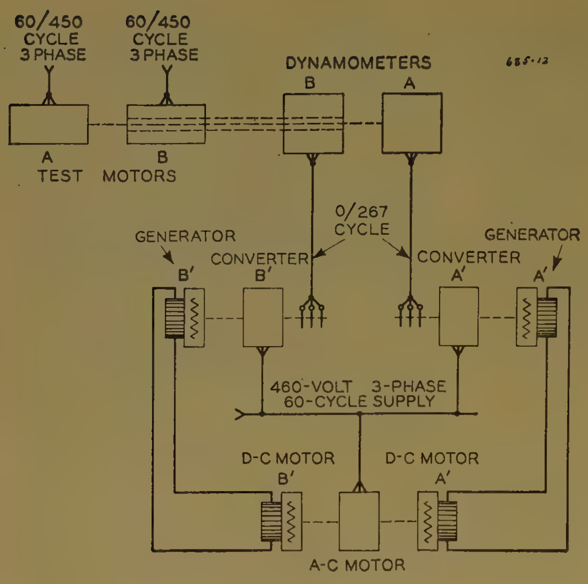


Figure 12. Schematic diagram of dynamometers and power absorption equipment connections

requirements, design studies indicated friction and windage losses to be in excess of the minimum torque values to be measured. Apart from any other considerations, this called for some means of supplying electric power to the stator of the dynamometer to compensate for the friction and windage losses. The substitution of induction-type machines in conjunction with adjustable frequency absorption equipment satisfies these requirements.

The induction-type dynamometer used consists of a squirrel cage induction motor with the stator cradled in oscillating trunnion bearings. Connections to the stator windings are made through mercury cups in order to minimize inconsistent interference which would reduce the precision of these instruments. These dynamometers act as induction generators, or in some cases (as explained later) as induction motors, and heat to be removed represents no more of a problem than in conventionally designed induction motors of equivalent rating. Ventilation is provided by constant speed blowers mounted on the cradled dynamometers, and connections are carried through mercury cups.

Reference to Figure 10 shows, by the shaded areas, the manner in which three cradled induction dynamometers were selected to cover the envelope of horsepower speed requirements for this installation.

Dynamometer A has a 100-horsepower absorption rating over a speed range of 3,500 rpm to 10,500 rpm. It is of solid shaft construction and when used, alone, will operate up to 16,000 rpm, at which speed the absorption capacity required is 30 horsepower. It also may be operated with dynamometer B either in tandem or for testing of contraturning model motor units. For either of the latter two mechanical arrangements, top speed is limited to 13,000 rpm, at which speed dynamometer A has the required capacity of 55 horsepower. Dynamometer B has a 100-horsepower absorption rating over a speed range of 3,500 rpm to 10,500 rpm. It is of hollow shaft construction, arranged for use with dynamometer A, either in tandem or for testing of contraturning model motor units. Under either of these conditions it may be operated up to a maximum speed of 13,000 rpm, at which speed it has the required capacity of 55 horsepower.

Figure 13. Operators' console in control room of Cornell variable density wind tunnel

Dynamometer C has a 35-horsepower absorption rating at 15,000 rpm and is capable of operating up to 27,500 rpm, at which speed it has more than the required capacity of 10 horsepower. It is of solid shaft construction. Figure 11 shows this dynamometer installation.

Reference to Figure 12 will show the manner in which the 100-horsepower dynamometers are connected electrically to feed energy back to the 60-cycle system. D-c motors A' and B' each are rated 60 kw (absorption), 250 volt, 1,800 rpm, and are driven by a 125-horsepower 440-volt 60-cycle squirrel cage induction motor. Induction frequency converters A' and B' each are rated 70-kw absorption and each connected to a 75-horsepower (input) 250-volt 500/1,800-rpm d-c generator. Each of the dynamometers A and B are of 2-pole construction and their synchronous speed frequencies corresponding to a speed range of 1,125 to 16,000 rpm are between 18 and 267 cycles. The converters A' and B' are designed to receive power on the slip rings over this frequency and, by varying the speed of d-c generators A' and B', deliver power from the stator to the 60-cycle system. The speed of generators $A'B'$ is changed between 0 and 500 rpm by armature voltage control. Above this value, armature voltage is maintained at 250 volts and speed is increased to 1,775 rpm by field weakening. At this point frequency on the converter slip rings is 267 cycles corresponding to 60 cycles at the stator terminals.

Consider the method of operating this system to calibrate test motor A on dynamometer A. The constant speed set, of

which motor A is a part, first is started by connecting the 125-horsepower a-c motor to the 440-volt 60-cycle line. Converter A' with stator winding connected to the 60-cycle line then is started from the d-c end in the same direction of rotation as its magnetic field. The speed of this set is raised by armature voltage control until the frequency on the slip rings of the converter is about five or ten cycles. At this point dynamometer A, with its model motor mechanically connected but electrically isolated, is connected electrically to frequency converter A'. The frequency applied to dynamometer A then is raised by first lowering the speed of the converter close to zero revolution per minute, then reversing it and raising the speed in the opposite direction. When the desired minimum test speed for the model motor is reached, its supply frequency will be matched, and the motor will be connected to its supply. Incidentally, this procedure for connecting the model motor to its supply also is followed for a model motor in the tunnel. In this case the model propeller with its motor is "windmilled" by the tunnel air stream, and the power supply must be matched to the motor speed before connecting it to the motor. In order to load the "test" motor, while its speed is held constant by its power supply, it is merely necessary to slightly lower the frequency at the terminals of dynamometer A. This is accomplished by vernier adjustment of the field of motor A.

The 35-horsepower dynamometer is operated in a similar manner by connecting it to a suitable adjustable frequency source.

Conclusion

These are some of the problems to be considered in the application of electric equipment for an installation of this magnitude. There are, of course, many others. However, this indicates to some degree their nature and complexity. More than that, it demonstrates how important is the correct selection of electric equipment.

Other laboratory facilities will have comparable problems, but electrical engineers will find that all demand the same approach. A careful analysis must be made, a practical, thorough, and detailed study of the performance expected by the scientists who are going to operate the new instrument.

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F-3 Lead Alloy—An Improved Cable Sheathing

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OPERATING EXPERIENCES have demonstrated that the attainment of good service reliability of most electric power cables has become substantially more a function of the characteristics of the sheath than of the insulation. Modern underground installations of impregnated-paper-insulated, lead-covered power cable include

1. Solid-type cable with
(a). Compound-filled joints without reservoirs.
(b). Oil-filled joints with reservoirs.

2. Oil-filled cable operated at 10- to 15-pound gauge pressure.

3. Gas-filled (or pressure) cable operated at
(a). Low pressure (10- to 15-pound gauge).
(b). Medium pressure (24- to 40-pound gauge).
(c). High pressure (150- to 300-pound gauge).

Aerial installations, in which the cable is suspended by rings to a messenger strung between poles, consist generally of the same basic cable designs.

Depending upon the type of cable and installation, modern requirements for lead sheath include some or all of the following properties in addition to the time-honored criteria of strength, ductility, hardness, "cleanliness," and concentricity:

1. Resistance to slow bending fatigue resulting from movement in the manholes caused by expansion and contraction during daily load cycles.

2. Resistance to creep (expansion) at low internal pressures (10- to 30-pound gauge).

3. Resistance to burst caused by high internal pressures developed at the bottom of slopes and vertical risers, in pressure-type cable systems, and in solid-type cable systems having reservoirs of oil, connected to the joints.

4. Resistance to abrasion caused by installation and movement in ducts.

5. Resistance to vibration fatigue caused by traffic on any type of installation and by wind on aerial cable.

6. Stability from age-hardening or self-annealing so that the foregoing properties are retained throughout the life of the cable and so that the cable can be withdrawn from ducts and reinstalled without cracking the sheath.

7. Workability, which facilitates melting and extruding as sheath with sound welds and uniform composition.

8. Corrosion resistance.

Some commercial grades of lead have given satisfactory service in a great many installations, especially up to about 20 years ago. The sheaths have become increasingly the limitation[1-5] on the serviceability of power cable with the advent of the following:

1. Decrease in viscosity of impregnating compound.

2. Use of oil reservoirs connected to joints.

3. Improvement in thoroughness of insulation impregnation.

4. Increase in maximum allowable temperatures for normal operation.

5. Establishment of still higher temperatures for emergency operation.

Item 5 was the last to occur, and its use has accentuated the need for an improved cable sheathing.

Numerous lead alloys have been developed to overcome these difficulties, but each, it seems, is lacking in one or more important property.

Antimony alloy (0.75 to 1.0 per cent) has poor creep resistance[6] and undergoes change in properties with age,[7] including loss in ductility.

Tin alloy (two per cent) has relatively poor creep resistance[6] and about the same or slightly better resistance to vibration fatigue than the best commercial grades of lead (refer to Figure 13).

Cadmium-antimony and cadmium-tin alloys, known as BNF alloys, offer extrusion difficulties, age-harden, and have poor creep resistance.[8]

Calcium alloys (0.025 and 0.031 per cent) are difficult to extrude (very drossy) and undergo slow and serious age hardening, which has caused troubles to develop in a few years of service because of brittleness and early cracking in the manholes.[9]

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Tellurium alloy (0.05 to 0.10 per cent) tends to segregate in the welds during extrusion and, according to unpublished data, has poor creep resistance.

An Improved Lead Alloy

To satisfy the need for a lead alloy having the desired improved properties, the Anaconda organization instituted a research program to develop a suitable alloy. The best material developed in this program is known as F-3 lead alloy.[10] It is composed of arsenical lead containing small amounts of tin and bismuth. This new material when suitably hardened by heat treatment also has physical properties which may meet the requirements set for medium-pressure gas-filled cable.[11] The sheaths are characterized by their strong tough welds, outstanding resistance to bending fatigue, excellent creep resistance, and bursting strength.

Chemical Composition

The chemical composition used for the F-3 alloy pig metal is

Arsenic (As)	0.15 per cent
Tin (Sn)	0.10 per cent
Bismuth (Bi)	0.10 per cent
Copper (Cu)	Minimum
Antimony (Sb)	Minimum
Lead (Pb)	99.65 per cent

This alloy is extruded in conventional equipment with the same technique employed in handling commercial grades of lead. It is a "clean" alloy not subject to excessive drossing associated with some other lead alloys. This feature results in strong welds and facilitates the making of wiped joints. Beginning early in 1945, a number of large orders of paper-lead cable with F-3 alloy sheath have been produced, shipped, and installed without difficulty.

Bending Resistance

The current carried by a power cable usually varies in a daily load cycle which causes a daily variation in temperature. This variation produces longitudinal expansion and contraction of the cable which accumulates in the manholes where expansion loops are provided. The accompanying daily bending of the cable in the manholes has, in some cases, broken the sheaths. Repairs have not proved very satisfactory, and in many cases replacement of lengths of perhaps 300 to 500 feet of cable has been necessary. For many installations, current-carrying capacity and life of the cable are determined more by the ability of the sheath to withstand such slow bending than by the ability of the insulation to withstand the voltage.[3,5]

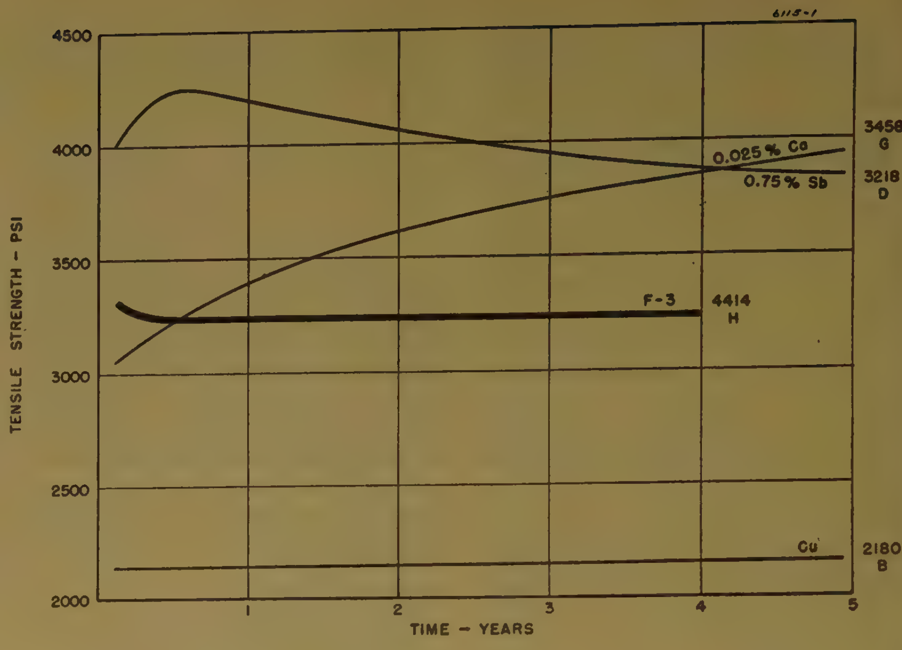


Figure 1. Age hardening

Room temperature

Tests have been conducted on the bending of various types of cable at room temperature and elevated sheath temperature. The results obtained on accelerated cyclic bending, 0.5-inch cable movement at each duct mouth, indicate that the life of F-3 sheath is between three and four times the life of copper-bearing lead sheaths on the same kind of cable.

Abrasion Resistance

A large majority of underground cables in this country are pulled into ducts. In recent years increasingly longer lengths and larger and heavier cables have been used. In some cases as much as seven tons or more of cable have been installed in one pull. Irregular ties or pebbles in the ducts cause scoring of the sheath. Also, it is subject to bruises in handling and working around it. After it is installed, expansion and contraction with daily changes in temperatures cause continual movement during the life in service.

In aerial cable, abrasion resistance is needed, not only to withstand the effects of installation and of cable movement caused by expansion and contraction, but also to minimize cutting of the sheaths at the rings which support the cable.

The most suitable indication of abrasion resistance is the hardness test, although it is recognized that some factors other than those involved in such tests are influential. The relative hardnesses of various lead and lead alloy materials are shown in Table I.

These hardness data apply to extruded material which has aged at room temperature for about three to eight months. Generally, lead alloys grow appreciably harder with age, but the hardness of F-3 alloy remains about the same.

While hardness is desirable to obtain abrasion resistance, the harder materials usually lack ductility and break under tensile stress with very small elongation. Age-hardening materials lose generally in ductility with age, becoming in some cases too brittle for use. The new F-3 alloy has both hardness and good ductility which is retained with age, as is shown later.

Long-Time Aging

The F-3 alloy does not harden with age, as calcium, antimony, and to some extent tin alloys do. Instead of an increase in tensile strength, a slight decrease occurs, as is shown in Figure 1. These data were obtained from tests of water-quenched tubes which were aged and tested at room temperature. Similar results were obtained in tests of sheath removed from cable.

Materials which harden appreciably with age lose ductility, a fact which is

Table I. Hardness of Lead and Alloys

Room Temperature

Sample Number	Material	Brinell Hardness Number
2,264	Desilverized lead—A	3.9
2,180	Copper lead	4.5
2,296	Chemical lead	4.5
3,458	0.025 per cent calcium	8.1
4,414	F-3 alloy	8.4
4,212	2 per cent tin	9.0
1,228	0.031 per cent calcium	10.4
3,218	0.75 per cent antimony	10.4

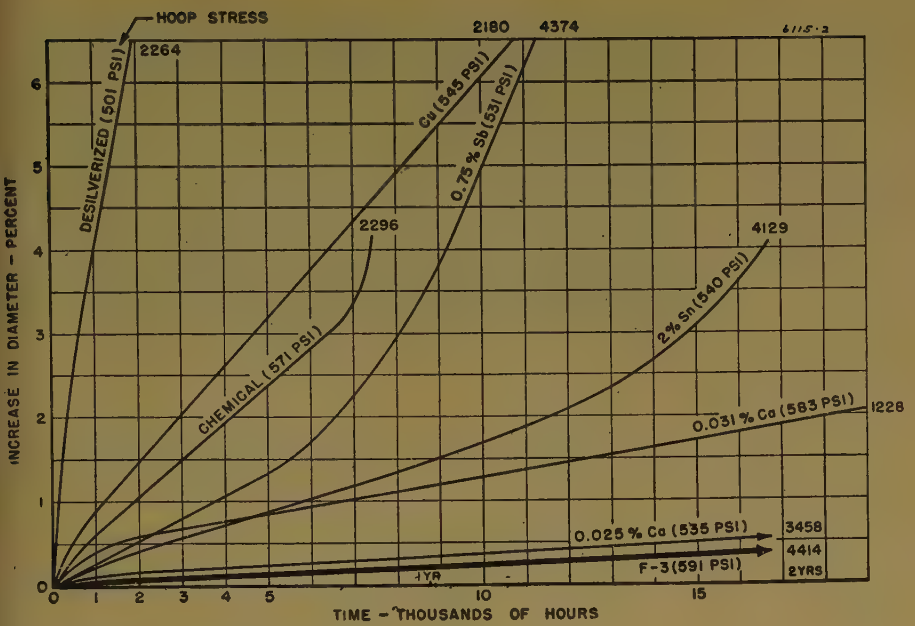


Figure 2. Creep of tubes

61-pound gauge, room temperature

shown by decreasing amounts of elongation to fracture in tensile tests of strips removed from cable sheaths. F-3 alloy has given an elongation to fracture of about 23 to 30 per cent throughout the four or five years during which tests have been made. Further information on this characteristic is given under "Fracture and Ductility."

Creep Resistance

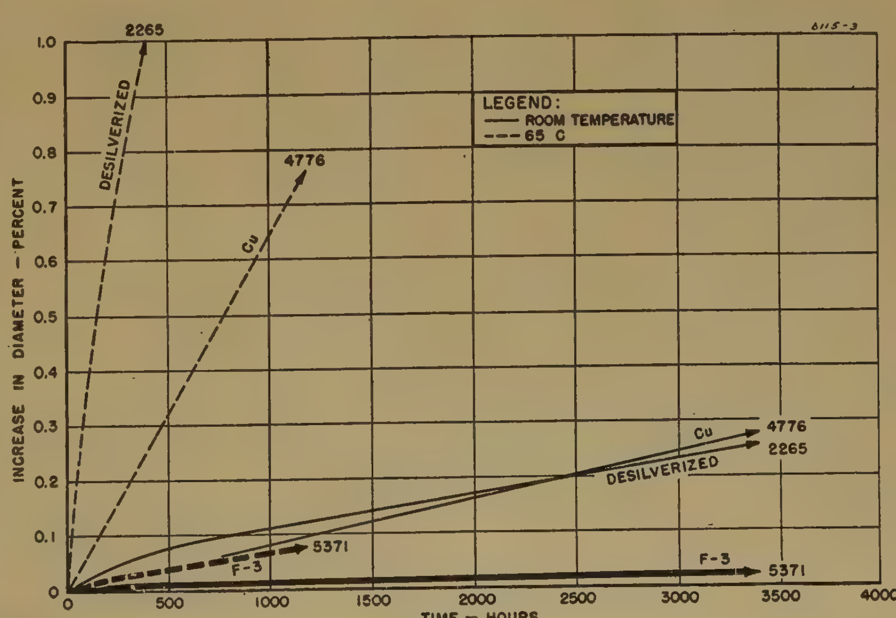
Resistance to creep under tensile stress is especially beneficial for sheaths on oil-filled and gas-filled cables and on solid-type cable having a hydrostatic head of oil. In all of these the continuous internal pressure tends to expand the sheath. Under relatively low stress for periods of years, the safe expansion of commercially pure lead sheaths is much less than the 30 to 50 per cent obtained to fracture in short-time tensile tests. It is of the order of three to five per cent. For most of the alloys the elongation that can be expected without fracture is less than that for lead in both short-time and long-time tests.

In solid-type cable having no additional oil available, the sheath is expanded by bending during installation and by thermal expansion of the conductors and insulation within the sheath. Since lead has little elasticity, a large part of the expansion remains as permanent distension of the sheath, leaving a gap on the inside. Void spaces thus are produced which are likely to appear in the insulation. Such voids are detrimental to the electric strength of the insulation in high-voltage cables. Thus good creep resistance is beneficial in obtaining long life for both the sheath and the insulation.

Tube or sheath specimens 15 inches long and without cable inside were sealed with compression fittings and subjected to constant internal oil pressures.[12] The creep (that is, increase in tube diameter) was measured by a special type Al-SR4 strain gauge[13] over a period of from one to five years. This method was found capable of measuring changes in diameter in the order of 0.002 per cent. Test lines were operated at gauge pressures of 61, 42, and 35 pounds at room temperature and at 35 pounds at 65 degrees centigrade.

Figure 3. Creep of tubes

35-pound gauge



Typical data, selected from the 105 samples tested, are shown in Figures 2 and 3 in diameter with time. From Figure 4 in which the data are shown as creep rates, it is evident that F-3 alloy has a creep resistance at room temperature about 45 times that of copper-lead and 25 times that of chemical lead at a stress of about 500 to 600 pounds per square inch and at least equal to any heretofore known lead-alloy sheath material. The curves in Figure 3 for creep at 35 pounds (approximately 300 pounds per square inch hoop stress) indicate that F-3 alloy has about 30 times the creep resistance of copper-lead at room temperature and about ten times that of copper-lead at 65 degrees centigrade.

Fracture and Ductility

Bursting Strength of Tubes

Tube specimens 15 inches long were tested at various stresses and the time required for failure obtained as described by Bassett and Snyder.[12]

Figure 5 shows the data obtained for F-3 alloy and other materials. (This figure is essentially the same as Figure 7 of reference 12 with a few deletions and the addition of new data.)

Data from fracture tests of lead show customarily that at low stresses, such as 400 and 600 pounds per square inch, lead alloys do not have as much advantage over common lead and copper-lead as their tensile strengths would indicate. At higher stresses, such as 1,000 and 1,500 pounds per square inch, the alloys, of course, are better. Figure 5 shows that in these tests at room temperature, F-3 alloy has bursting characteristics different from other alloys. It seems to retain its strength ratio to copper-lead equally at all stresses.

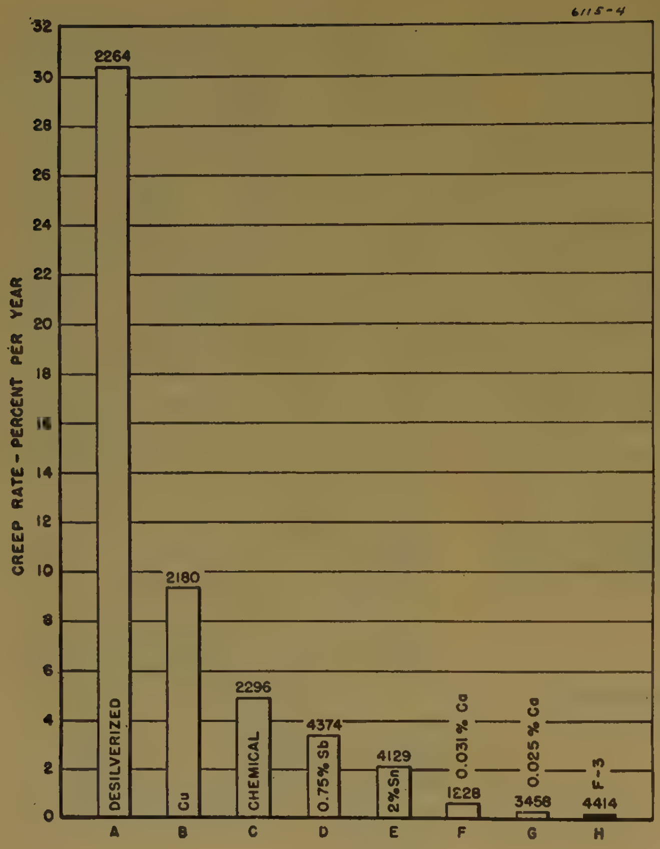


Figure 4. Creep rate of tubes

61-pound gauge, room temperature

prising the total footage from one to three complete lead press charges, were conducted to locate the weakest part of the length. These samples were sealed on the ends with compression fittings and subjected to continuous 85-pound air pressure until failure. After the initial failure, ten feet of pipe from the finish of the charge was resealed on the ends and the test continued until final failure.

Table II summarizes the data thus obtained.

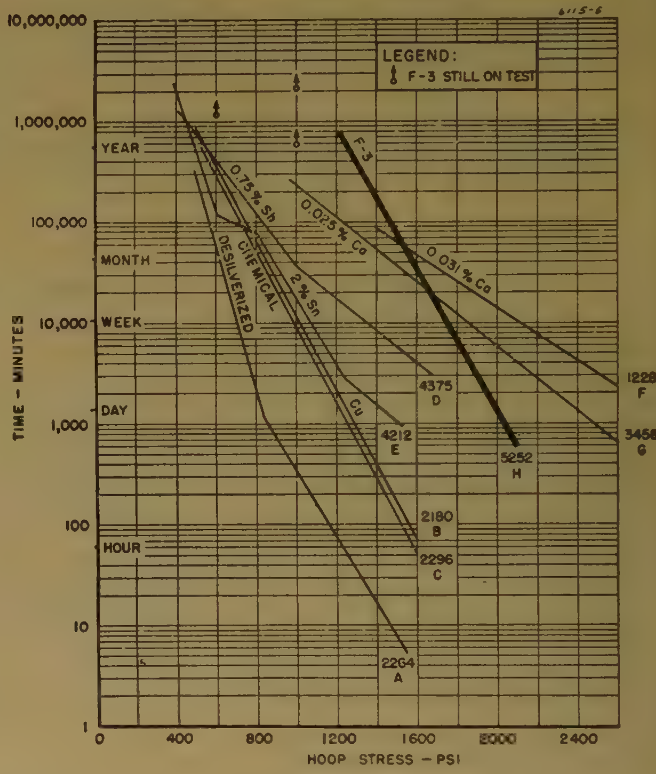


Figure 5. Bursting strength of tubes

Room temperature

This behavior leads to another point, that of the type or appearance of the failure. It has been observed that when lead fails under static long-time load at stresses up to approximately 30 per cent of ultimate strength (too low to cause appreciable plastic flow), the fractures are intercrystalline.[12,14,15] This type of failure is referred to as "stress cracking." The types of fracture produced by burst tests for commercial grades of lead and some alloys were described by Bassett and Snyder.

Figures 6, 7, and 8 show the type of failure obtained with F-3, calcium, and tin alloys. Other illustrations appear in reference 12.

No sample of F-3 alloy has failed in these bursting tests at stresses of 1,000 pounds per square inch or less, even though specimens have been on test for over four years. Specimens tested at 1,250 pounds per square inch have failed without stress cracking.

Data presented in Figure 9 indicate that F-3 alloy retains its good ductility in low-stress long-time tests as well as in short-time tests.

Bursting Strength of Long Lengths

Burst tests on long lengths (150 to 300 feet) of F-3 alloy pipe, each length com-

Table II. Bursting Strength of Full-Charge Lengths of Pipe

Approximately 850 Pounds Per Square Inch, Room Temperature

Material	Approximate Time to Failure, Months: Weakest Point of Full Length	Approximate Time to Failure, Months: Uniform 10 Feet at End of Charge	Location of Weakest Point of Full Length
Copper-lead	1	3	Charge weld
Tin alloy	1/2	3	Charge weld
Antimony alloy	1	9	Charge weld or where deformed by mechanical damage
Calcium alloy	1/2	24	Charge weld or where deformed by mechanical damage
F-3 alloy	12	No failure after four years	Adjacent to press-stop mark

Stress–Strain

Nonferrous metals including lead do not behave elastically; that is, they do not follow Hooke's law up to a certain stress point, the yield point. However, they do exhibit a certain degree of elasticity on short-duration tests. Elasticity, if present in a cable sheath, would be of some practical importance in producing some recovery in the radial expansion of the sheath after the initial high pressure developed during the start of a load cycle.

Test specimens 15 inches long like those used for burst tests were fitted with SR-4 strain gauges and subjected to repeated pressure cycles of successively increasing magnitude. The strain and recovery during each test cycle were noted and plotted in Figures 10 and 11. Each point on the curves represents about one minute elapsed time. The figures show that copper-lead undergoes rapid strain at stresses above about 500 pounds per square inch hoop

Figure 6. Burst failure of F-3 alloy

Ductile break (magnified four times), sample 5,252
1,250 pounds per square inch, 10,400 hours, 18.4 per cent expansion

stress, whereas F-3 alloy withstood about three times this stress to produce the same amount of permanent set.

Vibration Fatigue

Flattened-strip specimens were machined with reduced sections and carefully calibrated for stress calculations. These were flexed in a vibrating-reed Bell-Laboratories-type machine at 750 cycles per minute, the deflection being varied to obtain different stresses in duplicate specimens.

Figure 12 shows the data as stress versus number of cycles (S/N) curves. The resistance to vibration fatigue of F-3 alloy is much superior to that of copper-lead and also superior to that of antimony and tin alloys which have been used commonly for sheaths on overhead power cables.

Corrosion

Laboratory corrosion tests indicate that F-3 alloy has resistance to ground water and acidic fluids equivalent to that of most high-purity leads.[16] It has been observed that F-3 alloy sheaths are brighter and remain brighter in ordinary atmospheric exposure than copper-lead.

In general, it has been found that where corrosion is experienced, mitigating measures are required regardless of the chemical composition of the sheath.

Joint Sleeves

If the joints on impregnated-paper-insulated cable are completely filled with compound and sealed without any provision for thermal expansion, then internal pressures developed in the cable during load cycles will appear also in the joints. Joint sleeves are usually considerably larger in diameter than the cable sheaths, and hence for given internal pressures, the hoop

Figure 7 (above). Burst failure of 0.025 per cent calcium alloy

Intercrystalline break (magnified four times), sample 1,867
1,500 pounds per square inch, 292 hours, 1.7 per cent expansion

Figure 8. Burst failure of two per cent tin alloy

Intercrystalline break (magnified four times), sample 4,212
1,000 pounds per square inch, 216 hours, 1.8 per cent expansion

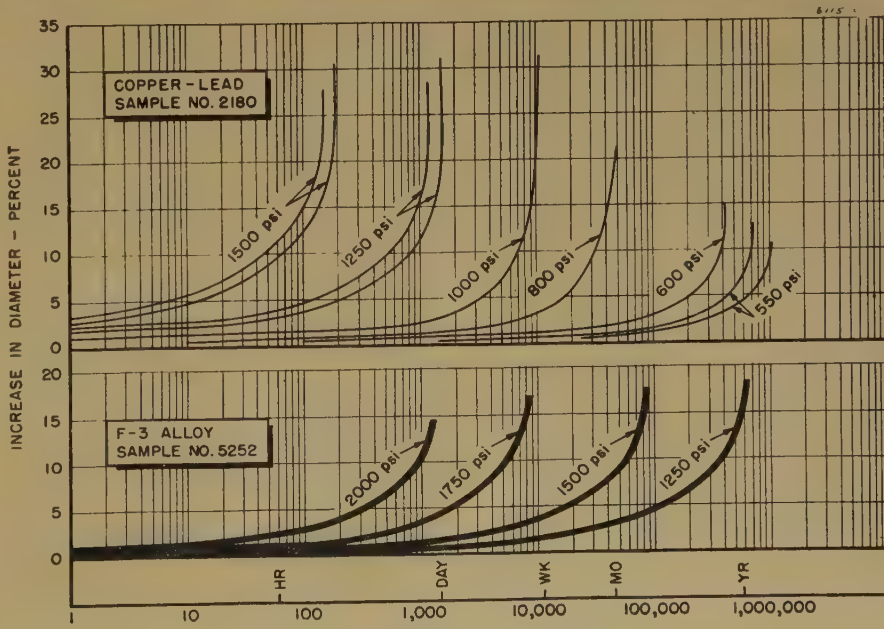


Figure 9. Ductility of tubes (room temperature)

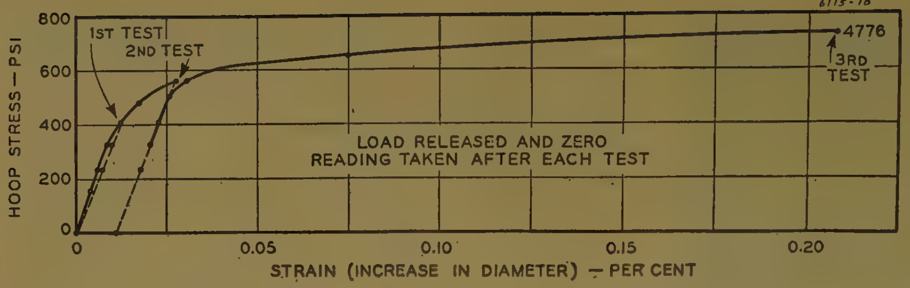


Figure 10. Stress–strain, copper-lead (tubes, room temperature)

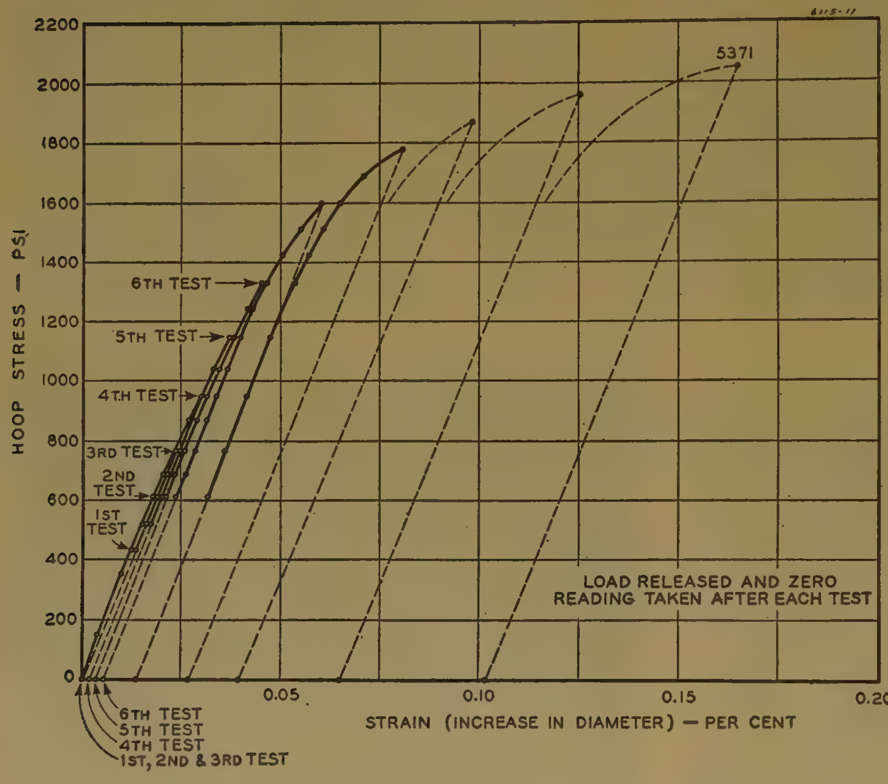


Figure 11. Stress–strain, F-3 alloy (tubes, room temperature)

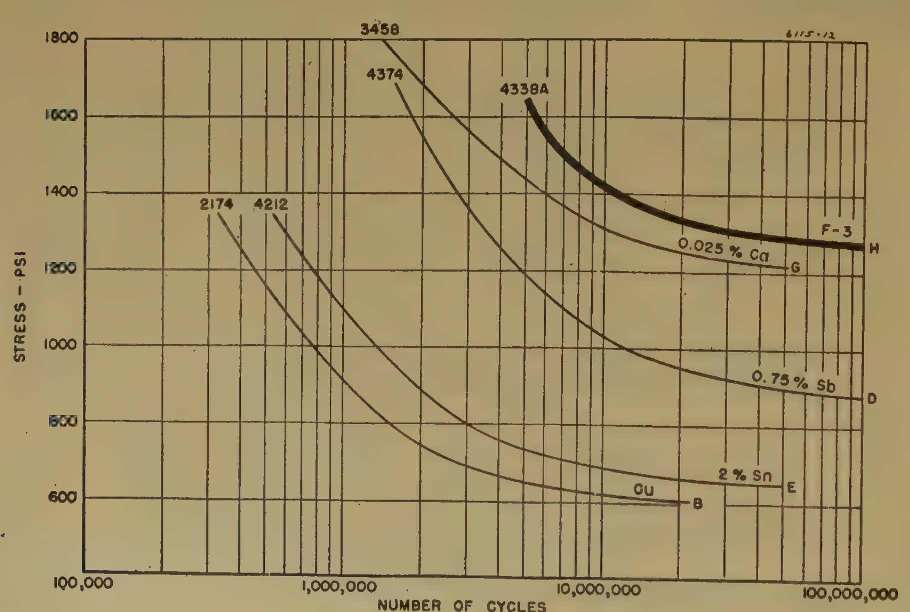


Figure 12. Vibration fatigue of flat strips
750 cycles per minute, room temperature

stresses in the sleeves are usually considerably greater than those in the sheaths unless the sleeves are made appreciably thicker.

With F-3 alloy for the cable sheath, having somewhat greater strength and considerably lower creep rate than lead sheaths, higher internal pressures can be expected in the sheaths and joint sleeves. It will become desirable, therefore, either to use material stronger than lead for the joint sleeves or to use some method of reinforcing the joint casing. Thus it may be quite desirable to have the joint sleeves made of F-3 alloy as well as the cable sheaths. Test joints have been made from extruded F-3 alloy sleeves and no difficulty was experienced in beating the ends or in wiping the joints.

Advantages of Improved Cable Sheathing

For some underground cables, the life in service is limited by the ability of the sheath to withstand repeated bending in the manholes caused by cable movement. In such cases the use of F-3 alloy will permit either higher loading for equal life of the cable or longer life with equal loading. For many installations, changes in the system or other factors make it more economical to obtain greater usage of the investment rather than to obtain much greater life.

The lower creep rate of the F-3 alloy will permit the use of higher internal pressures. This is especially beneficial on either oil-filled or gas-filled cables. A limiting hoop stress of 125 pounds per square inch has been established for design purposes for copper-bearing lead sheaths. This limitation restricts the internal pressures for oil-filled cables with standard sheath thicknesses to about 14-pound gauge for 2 $^1/_2$-inch cable or 12-pound for 3 $^1/_2$-inch cable. To withstand higher internal pressures, lead sheaths must be reenforced with layers of paper and copper tapes which require a costly double-sheath construction.

With F-3 alloy sheath, the equivalent stress for a similar creep rate is about 200 to 300 pounds per square inch based on data for tube specimens. If the limit is tentatively set at 200 pounds per square inch then with standard thicknesses the internal pressures can be increased safely to 22.5 pounds for 2 $^1/_2$-inch-thick cable or 19 pounds for 3 $^1/_2$-inch cable. By increasing the sheath thicknesses 40 or 50 mils, it becomes possible to use single sheaths on pressure-type cables with internal pressures of 25- or 30-pound gauge. This would avoid the use of double-sheath construc-

tion on, for example, some medium-pressure gas-filled cables.

Since cable sheathing must withstand the working and wear involved in installation which produce gouging and scoring of the outer surface, the thicknesses for most solid-type cables should not be reduced much from the present standard thicknesses. The sheath must be sufficiently thick to have long life in spite of such damages. The present standard thicknesses for lead have been evolved from long experience and are known to be adequate for normal conditions of use. At least until more experience is obtained in the use of F-3 alloy, the same thicknesses as for lead are recommended. For special conditions, where in the past extra-thick sheaths of ordinary lead have been required, the use of F-3 alloy will warrant a reduction.

Conclusions

1. A new cable sheath alloy of arsenical lead (F-3 alloy) has been produced. It has been successfully applied in manufacture to cable which has been installed in commercial lines.

2. F-3 alloy has been shown to have much better properties for power cable sheaths than any of the materials previously tried.

3. With this new type of sheath, the larger cables can carry higher loads without shortened life caused by cracking of the sheaths in daily bending in the manholes.

4. The limiting internal pressures will be appreciably higher than for lead sheaths, which may permit the use of single sheaths on some pressure-type cables where with lead the more costly double-sheath construction is required.

5. Since this alloy is harder than commercially pure lead, a reduction in troubles in service caused by mechanical damages can be expected.

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Electrical Accuracy of Selsyn Generator—Control Transformer System

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THE military requirements for high accuracy position control systems have placed emphasis on the need for a voltage signal source which can indicate electrically a mechanical shaft position. In addition this signal source must have high accuracy, marked dependability, and the capability of continuous mechanical rotation. A modification of the conventional Selsyn generator-motor system[1] has been developed and used extensively to fill this need.

This system,[2] shown schematically in Figure 1, consists of a generator in which a single phase a-c field is established with its space orientation determined by the position of the rotor, electrically connected to a control transformer in the rotor winding of which is induced a similar alternating voltage whose magnitude and polarity depend on the relative position of this rotor with respect to that of the generator. Ideally the control transformer rotor voltage versus the difference in the rotor angular position characteristic is a sinusoidal one, as is shown in Figure 2.

In effect, the presence of the single-phase magnetic field in the generator rotor induces voltages in the three stator windings of the generator. These voltages impressed on the three stator windings of the control transformer produce currents there which establish a single-phase magnetic field in the air gap of the same space position as that of the generator rotor. Depending on the actual orientation of the control transformer rotor with respect to the generator rotor, the voltage induced in the rotor winding may vary from a maximum of one polarity through zero to a maximum of the other polarity. If the control transformer rotor is positioned so as to maintain zero voltage across its terminals, it then maintains a fixed angular relationship to the generator rotor. Hence, it is possible to position the control transformer shaft properly with respect to the generator position by maintaining zero voltage on the control transformer terminals.

Although this system inherently has a high accuracy of the order of 0.3 degree, it is not completely error-free. To design Selsyn systems which are more accurate, it is necessary therefore to determine the effect of various factors which can contribute to electrical errors. The conditions under which errors are to be considered here are

1. Stationary orientation of both machines. This condition also explains the predominate error phenomenon from standstill up to speeds of three or four per cent of synchronous speed.

2. Constant velocity for both machines with a fixed angular separation between them. This condition considers the effect of "speed" voltages caused by rotation at speeds of from three or four per cent of synchronous speed on up to and above synchronous speed.

Conclusions

From calculations based on the analysis presented here, as well as other analyses, and from test measurements, the following conclusions are evident:

1. Static errors are of two general types, those which are predominately
(*a*) in time phase with the normal position error, or
(*b*) in time quadrature (out of phase) with the normal position error.

2. The in-phase errors are caused by unbalanced reactance either in the stator windings or the interconnecting lines or in the mutual reactance between rotor and stator of either machine. For this condition, zero magnitude control transformer voltage can be obtained, but not with the control transformer rotor 90 degrees from the generator rotor.

3. The out-of-phase errors are caused by unbalanced resistances in the stator windings or lines between generator and control transformer. For this condition there is no one control transformer rotor position where zero magnitude voltage is obtained.

4. Errors present at standstill are minimized by proper electrical design and care in manufacture. By judicious choice of the pitch, distribution, and skewing of windings and by properly shaping the pole face of the rotors, the fifth and seventh harmonics of mutual flux between rotor and stator of the two machines are eliminated and the in-phase errors thereby are reduced. Care in manufacturing insures that there is a minimum of phase unbalance of resistance or reactance, thus tending to eliminate the errors noted in items 2 and 3.

5. Constant velocity errors tend to be a combination of the in-phase and out-of-phase type. At low speeds the dominant effect is to shift the rotor position at which zero voltage is obtained, there still remaining a small out-of-phase component. At high speed the dominant effect is the introduction of an out-of-phase voltage, the so-called speed voltage.

6. The constant velocity errors are dependent on the ratio of the actual speed of rotation to the synchronous speed for the given excitation frequency. Hence, the use of higher excitation frequencies such as 400 cycles instead of 60 cycles would tend to minimize these effects.

Analysis

Accuracy at Standstill

The generator-control transformer system at standstill appears as three inductive coupled circuits as shown in Figure 3. The steady-state a-c equations of voltages for these circuits are

$$[r_{ag}+r_{ac}+j(X_{ag}+X_{ac})]i_a+ [r_{abg}+r_{abc}+j(X_{abg}+X_{abc})]ib+ jX_{MDag}I_D=0 \quad (1)$$

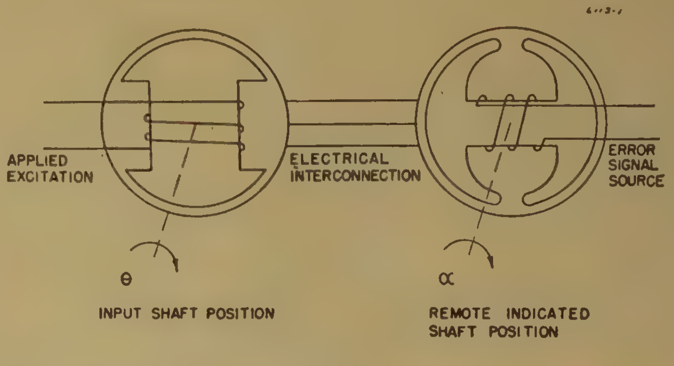

Figure 1. Schematic diagram of Selsyn generator-control transformer system

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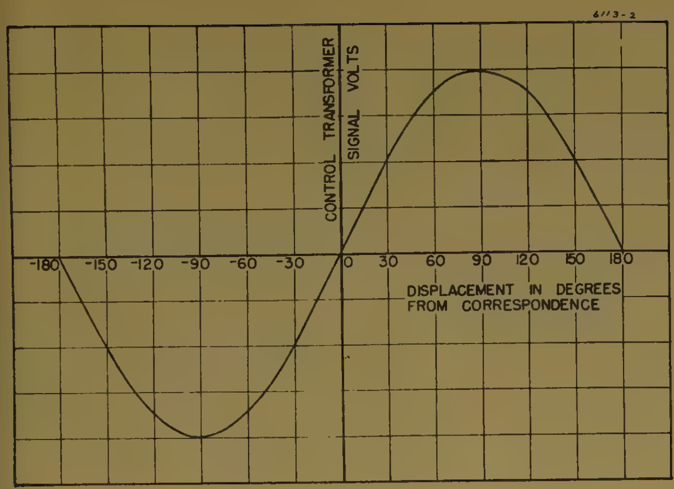


Figure 2. Ideal transformer voltage output versus angular displacement

$$[r_{abg}+r_{abc}+j(X_{abg}+X_{abc})]i_a+[r_{bg}+r_{bc}+j(X_{bg}+X_{bc})]ib+jX_{MDbg}I_D=0 \quad (2)$$

$$jX_{MDag}i_a+jX_{MDbg}i_b+(R_D+jX_{Dg})I_D=E_g \quad (3)$$

The various symbols used are defined in the nomenclature and are consistent with the synchronous machine representation of Doherty and Nickle,[3] Prentice,[4] and Park.[5] It will be noted that the nomenclature provides direct relationships between line-to-line reactances, easily obtainable from test, and the direct and quadrature reactances, which are very convenient for analytical work and are used extensively in the references cited above.

Figure 3. Elementary diagram of Selsyn generator-control system at standstill

The mutal reactances X_{Mbag} and X_{MDbg} here considered include only the lower order harmonics, that is, up to the eleventh. Only odd harmonics are present because of the salient pole construction, and the third and ninth components are absent because of the 3-phase stator winding arrangement. Higher harmonics could well be included in a similar fashion. The control transformer is effectively a round-rotor machine and is represented as such.

From equations 1–3, the vector currents i_a and i_b which have a magnitude and time phase can be determined. The voltage which exists in the control transformer rotor is dependent on these currents and α, the angular position of the control transformer rotor. Thus,

$$E_c=jX_{M1c}[i_a \cos (\alpha+30)+i_b \cos (\alpha-30)] \quad (4)$$

The Selsyn control transformer indicating system depends on there being a fixed angular displacement between the generator rotor position, θ, and the control transformer position of zero voltage. This displacement is normally 90 degrees hence the position error ϵp, can be defined as

$$\epsilon p=\theta-(\alpha+90) \quad (5)$$

where θ is the independent variable used in equations 1–3 and α is calculated from the condition $E_c=0$ in equation 4.

Referring to equation 4, zero control transformer volts is obtained when

$$i_a \cos (\alpha+30)+i_b \cos (\alpha-30)=0 \quad (6)$$

Assuming that the currents i_a and i_b from equations 1–3 may not be in time phase, and taking the instant when i_a is zero as a time reference,

$$i_a=I_a \sin \omega t \quad (7)$$

$$i_b=I_b \sin (\omega t+\beta) \quad (8)$$

where I_a and I_b are the current magnitudes, and β is the phase angle between them.

Combining equation 6–8 and separating the resultant expression into sin ωt and cos ωt components,

$$\sin \omega t[I_a \cos (\alpha+30)+I_b \cos \beta \times \cos (\alpha-30)]+\cos \omega t[I_b \sin \beta \cos (\alpha-30)]=0 \quad (9)$$

For equation 9 to be equal to zero for all values of time, the coefficients of the sin ωt and cos ωt terms both must be zero. With $\beta=0$, the cost ωt coefficient goes to zero, and the coefficient of sin ωt gives the following defining condition for α at zero control transformer voltage:

$$\tan \alpha=-1.732\left(\frac{I_b+I_a}{I_b-I_a}\right) \quad (10)$$

Referring to equation 9 and noting that in the usual case the time phase difference β is small, one realizes that by positioning α in the location defined by equation 10, the sin ωt component of voltage is made zero.

Hence there remains only a cos ωt component of voltage which, since β is small, is also small.

For the case where $\beta \neq 0$, the control

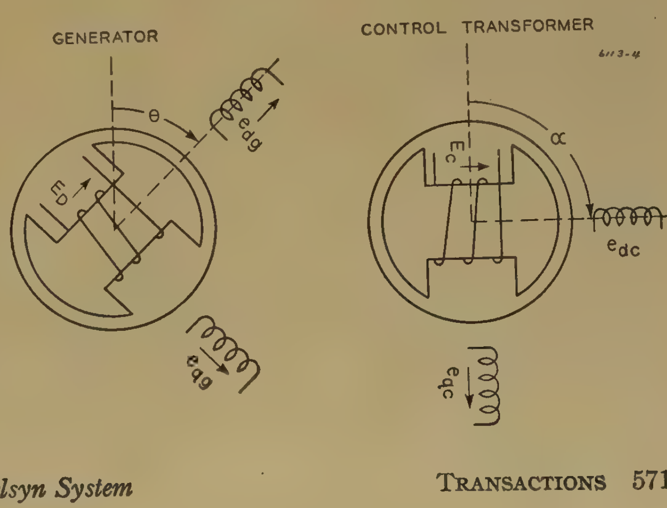


Figure 4. Two-reaction theory presentation of generator and control transformer

transformer error is then a voltage the magnitude of which is

$$e_u = I_b X_{M1c} \sin\beta \cos(\alpha+30) \qquad (11)$$

The foregoing analysis indicates the significant fact that there is a 90-degree difference in time phase of the errors e_p, where the voltages are of $\sin\omega t$ time phase, and the errors e_u, where the voltages are of $\cos\omega t$ time phase.

Through use of the analytical means developed in the preceding analysis, the effects of phase unbalance in such machine parameters as reactance, resistance, and mutual coupling have been investigated.

Data for generator:

$r_{ag}=10$ ohms	$E_g=115$ volts
$r_{abg}=5$ ohms	$\omega=377$ radians per second
$r_{bg}=10$ ohms	
$R_D=5$ ohms	$X_{M6g}=0$ except as noted on curves
$X_{dg}=120.5$ ohms	
$X_{qg}=39.5$ ohms	$X_{M7g}=0$ except as noted on curves
$X_{Dg}=110$ ohms	
$X_{M1g}=95$ ohms	

Data for control transformer:

$r_{ac}=100$ ohms	$X_{dc}=X_{qc}=1{,}200$ ohms
$r_{abc}=50$ ohms	
$r_{bc}=100$ ohms	$X_{M1c}=1{,}000$ ohms

Accuracy Under Constant Velocity Operation

In the treatment of the generator-control transformer system for this condition, it is premised that both machines have balanced windings. The effect of resistance or reactance phase unbalance is not considered, and only the fundamental component of mutual reactance between rotor and stator winding is included.

The following equations may be written for the generator machine in Figure 4, where the direct and quadrature axis voltage equations include transformer and speed voltage effects in the manner of Park.[5]

$$e_{dg} = r_{dg} i_{dg} + p\psi_{dg} - \psi_{qg} p\theta \qquad (12)$$

$$e_{qg} = r_{qg} i_{qg} + p\psi_{qg} + \psi_{dg} p\theta \qquad (13)$$

$$E_D = R_D I_D + p\psi_{Dg} \qquad (14)$$

Where the flux linkages are related to reactances and currents by the following equations

$$\psi_{dg} = X_{dg} i_{dg} - X_{M1g} I_D \qquad (15)$$

$$\psi_{qg} = X_{qg} i_{qg} \qquad (16)$$

$$\psi_{Dg} = X_{Dg} I_D - X_{M1g} i_{dg} \qquad (17)$$

Eliminating equation 14 and using equations 15–17, equations 12 and 13 may be written as

$$e_{dg} = (r_{dg} + pX'_{dg}) i_{dg} - pE'_D - X_{qg} i_{qg} p\theta \qquad (18)$$

$$e_{qg} = (r_{qg} + pX_{qg}) i_{qg} + X'_{dg} i_{dg} p\theta - E'_D p\theta \qquad (19)$$

Where the symbols X'_{dg} and E'_D are abbreviations of more complex quantities, namely,

$$X'_{dg} = X_{dg} - \frac{pX^2_{M1g}}{R_D + pX_{Dg}} \qquad (20)$$

$$E'_D = \frac{X_{M1g}}{\omega}\left[\frac{E_D}{R_D + pX_{Dg}}\right] \qquad (21)$$

For the control transformer, there is no

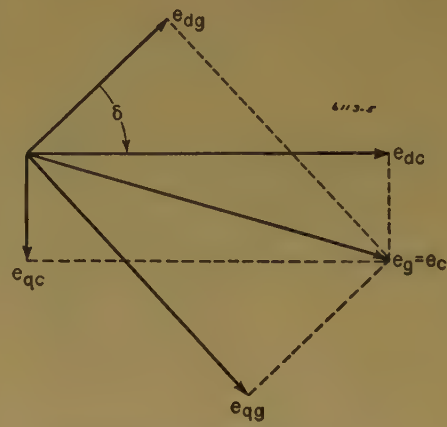


Figure 5. Relationship between *d* and *q* axes in the two machines

rotor current, so the equations for this machine are

$$e_{dc} = (r_{dc} + pX_{dc}) i_{dc} - X_{qc} i_{qc} p\alpha \qquad (22)$$

$$e_{qc} = (r_{qc} + pX_{qc}) i_{qc} + X_{dc} i_{dc} p\alpha \qquad (23)$$

In terms of the actual phase voltages, the terminal conditions for generator and control transformer are the same. The phase currents for generator and control transformer, however, are opposite in sign. From these relationships, one can relate the direct and quadrature components of voltages and currents for the generator in terms of the control transformer.

$$e_{dg} = e_{dc}\cos\delta - e_{qc}\sin\delta \qquad (24)$$

$$e_{qg} = e_{dc}\sin\delta + e_{qc}\cos\delta \qquad (25)$$

$$i_{dg} = -i_{dc}\cos\delta + i_{qc}\sin\delta \qquad (26)$$

$$i_{qg} = -i_{dc}\sin\delta - i_{qc}\cos\delta \qquad (27)$$

$$\delta = \alpha - \theta \qquad (28)$$

For continuous operation at a constant velocity,

$$p\theta = p\alpha = V\omega \qquad (29)$$

where

$$V = \frac{\text{actual speed}}{\text{synchronous speed at frequency } \omega}$$

Under steady-state conditions with excitation frequency ω, the operator p preceding the reactance terms is replaced by j. The $p\theta$ and $p\alpha$ velocities are handled as indicated in equation 29. Finally the instantaneous voltage E_D is replaced by the corresponding vector voltage,

$$E_D = \sqrt{2}E_g \sin\omega t = E_g\underline{|0} \qquad (30)$$

In vector terms, E'_D is equivalent to

$$E'_D = \frac{X_{M1g}}{\omega}\left[\frac{E_g\underline{|0}}{R_D + jX_{Dg}}\right] \qquad (31)$$

Combining the preceding equations and simplifying the results yield the following two expressions:

$$Ci_{dc} + Di_{qc} = j\omega E'_D \qquad (32)$$

$$Mi_{dc} + Ni_{qc} = +V\omega E'_D \qquad (33)$$

where C, D, M, and N are symbols which are defined as

$$C = -[r_{dg} + r_{dc} + j(X'_{dg} + X_{dc})] \times \cos\delta + V(X_{qg} + X_{dc})\sin\delta \qquad (34)$$

$$D = [r_{dg} + r_{qc} + j(X'_{dg} + X_{qc})] \times \sin\delta + V(X_{qg} + X_{qc})\cos\delta \qquad (35)$$

$$M = [r_{qg} + r_{dc} + j(X_{qg} + X_{dc})] \times \sin\delta + V[X'_{dg} + X_{dc}]\cos\delta \qquad (36)$$

$$N = [r_{qg} + r_{qc} + j(X_{qg} + X_{qc})] \times \cos\delta - V[X'_{dg} - X_{qc}]\sin\delta \qquad (37)$$

From equations 32 and 33

$$i_{dc} = \frac{Nj\omega E'_D - DV\omega E'_D}{CN - DM} \qquad (38)$$

and the actual control transformer output voltage is

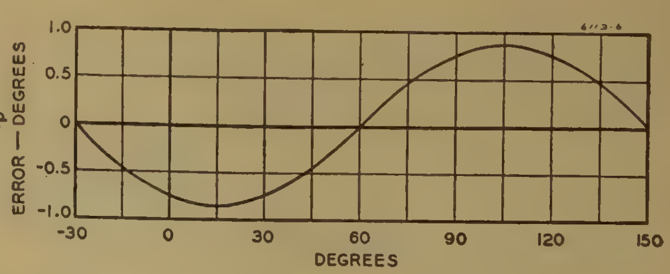
$$E_c = -jX_{M1c} i_{dc} \qquad (39)$$

Calculations were made using equations 31–39 for speeds ranging from 0 to synchronous speed, using the following data:

Generator:

$r_{dg} = r_{qg} = 29.7$ ohms $\qquad X_{Dg} = 208$ ohms

Figure 6. Control transformer error versus generator rotor position with unbalanced stator phase reactance $\dfrac{\Delta(X_{ag}+X_{ac})}{X_{ag}+X_{ac}} = 1$



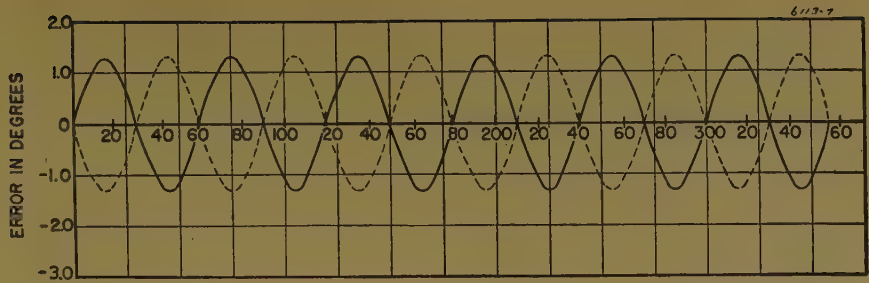


Figure 7. Control transformer error versus generator rotor position with harmonics of generator mutual reactance between rotor and stator

Solid line—$X_{MDag}=95 \cos(\theta+30)+2\cos(5\theta+30)$

Broken line—$X_{MDag}=95 \cos(\theta+30)+2\cos(7\theta+30)$

$X_{dg}=230$ ohms	$\lvert E_g\rvert=115$ volts
$X_{M1g}=204$ ohms	$\omega=377$ radians per second
$X_{qg}=62$ ohms	
$R_D=12.7$ ohms	$V=0$–1.0

Control transformer:

$r_{dc}=r_{qc}=686$ ohms
$X_{dc}=X_{qc}=3{,}320$ ohms
$X_{M1c}=1{,}860$ ohms

Results

The qualitative nature of the results has been verified by test and calculations. The quantitative results are not applicable to all machines but do indicate a performance representative of the more common type of generator-control transformer combination. In studying the effects of resistance and reactance unbalances, the magnitude of the quantities were chosen arbitrarily and do not reflect the usual values of machine constants.

The accuracy of Selsyn systems in use at present is remarkably high. As shown in Figure 8, position errors not in excess of 0.2 degree are obtainable. Although the generators and control transformers themselves are manufactured with resistance unbalances of less than ten ohms, an out-of-phase voltage of 0.25 volt, such as shown in Figure 9, may be obtained when these machines are used in system including switches.

Although the results are discussed in two separate sections, namely static accuracy and constant velocity accuracy, it should be noted that in actual operation errors due to both unbalance and velocity exist simultaneously and may combine in a number of ways to make up the total error. Hence, comparisons with test measurements under dynamic conditions tend to be difficult to analyze.

Static Accuracy

Since the impedance of the generator and control transformer is largely reactive, it follows that an unbalance in the phase values of reactance of either machine causes the in-phase current magnitudes to be altered from the values necessary for errorless operation. This shifts the flux orientation in the control transformer and requires that the rotor be moved slightly from its position of correspondence with the generator.

Specifically, unbalance in the stator phase values of reactance produces a space error which varies with the second harmonic of the generator rotor orientation (Figure 6). The maximum value of error for values of unbalance reactance up to about ten per cent is given by the expression

$$\epsilon p \max = 8^\circ \frac{\Delta(X_{ag}+X_{ac})}{X_{ag}+X_{ac}}$$

This relationship applies for machine constants given in the data in the section on static accuracy.

The presence of fifth and seventh harmonics in the mutual flux between rotor and stator of either machine is evidenced by the occurrence of a space error which varies as the sixth harmonic of the generator rotor orientation. As shown in Figure 7, the direction of the error is opposite for positive values of the fifth and seventh harmonics of mutual flux so that theoretically a balance can be obtained where one compensates for the other. The magnitude of the maximum error varies about linearly with the magnitude of the harmonic flux, thus

$$\epsilon p \max = +57.3^\circ \frac{X_{M5g}}{X_{M1g}} \text{ for + fifth harmonic}$$

$$\epsilon p \max = -57.3^\circ \frac{X_{M7g}}{X_{M1g}} \text{ for + seventh harmonic}$$

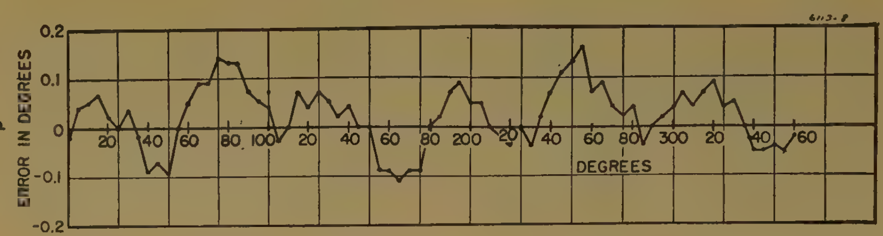


Figure 8. Typical generator-control transformer error versus generator rotor position

From test data

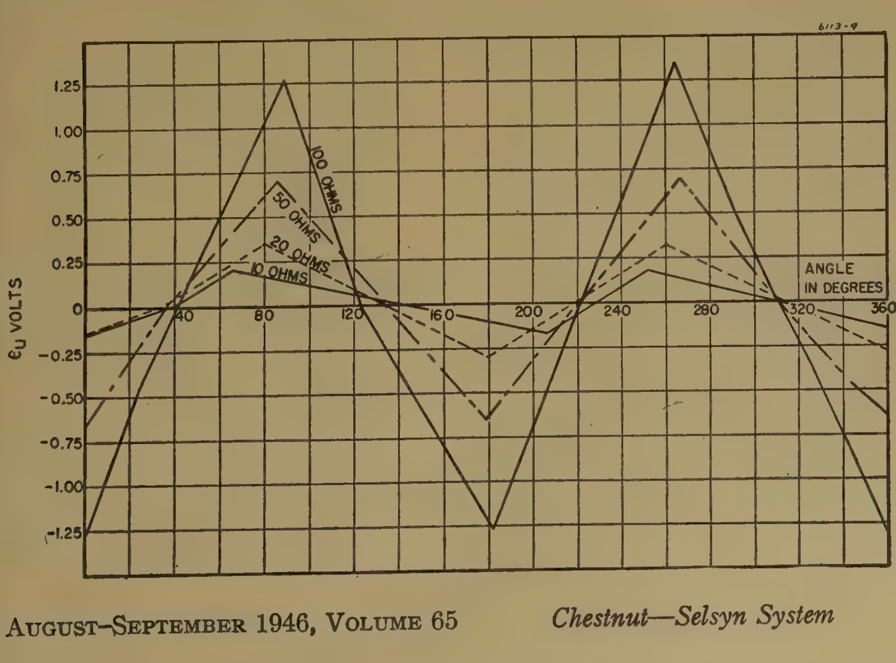


Figure 9. Control-transformer quadrature error voltage as a function of generator rotor angle

For various amounts of added resistance in one stator line

Points of maximum, minimum, and zero values of voltage are joined by straight lines

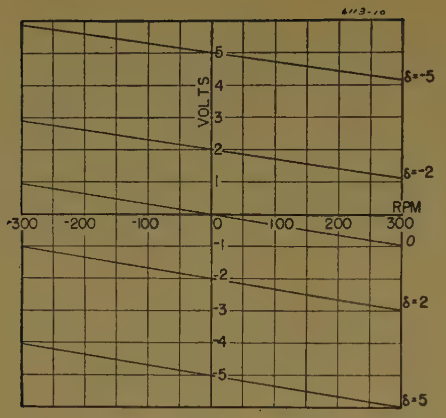


Figure 10. Calculated values of control transformer voltage (in phase component) versus speed

For generator-control transformer with fixed angular separation (δ)

Figure 7 verifies these approximate relations for the case of a machine with constants as given in the section on static accuracy.

In Figure 8 is shown a typical error curve measured on a generator-control transformer and indicating a pronounced sixth harmonic.

By modification of the generator rotor flux distribution to reduce the seventh harmonic component, the resultant error was reduced. The fifth harmonic of flux is absent from this type machine because of the stator slot winding arrangement.

The presence of unequal values of resistance in the different stator phases causes a shift in the time phase of the stator currents. The current magnitudes which are determined primarily by the phase reactances are relatively unchanged. Under this condition there is no control transformer rotor position where zero voltage exists at all instants of time. At zero ϵp error position, there exists a voltage induced in the control transformer rotor which is in time quadrature with the normal position error voltage. Turning the control transformer rotor from the zero space error position merely brings the voltage in phase with the normal signal voltage and greatly increases its magnitude. The polarity of the voltage depends on the direction in which the rotor is moved. The magnitude of the time quadrature component of voltage is about linearly proportional to the amount of unbalanced resistance as shown in Figure 9. This curve, obtained experimentally, verifies calculations which indicate a second harmonic variation of control transformer rotor minimum voltage as a function of generator rotor position.

Constant Velocity Accuracy

A generator-control transformer system which has zero error at standstill suffers a shift in the angular relationship between generator and control transformer rotor positions where zero control transformer rotor voltage is obtained when the two machines are running at the same speed. Stated conversely, if the angular relationship between the two rotors is maintained fixed and the two rotors are run at a number of constant speeds, the control transformer rotor voltage changes magnitude as a function of speed, the frequency of the output voltage being maintained constant throughout. Figure 10 shows the calculated values for the in-time-phase component of rotor voltage as a function of speed for a number of fixed values of relative rotor displacement. These curves compare fairly well with the experimentally measured curves shown on Figure 11, which show the smoothed "average" voltage-speed effect for the same type machines as were considered in the calculations. The presence of errors caused by irregularities of phase resistance or reactance introduces spurious errors which tend to obscure measurements of the speed voltage effect at these low voltages where the effect amounts to about 1 volt (=1 degree) per 320 rpm. Actually the discrepancy between calculations and test is well within the limit of experimental accuracy.

In Figure 12 is shown the in-phase and out-of-phase components of control transformer rotor voltage as a function of speed up to synchronous speed for a number of fixed values of angular separation. The applied voltage to the generator rotor is taken as the in-phase voltage reference. A time phase shift of about 15 degrees at 60 cycles per second takes place in this generator-control transformer system at standstill. In Figure 13 the control transformer output voltage at standstill is chosen as the reference for the in-phase voltage, otherwise the data are the same as on the preceding figure. From these curves it can be seen that the effect of high speed rotation is primarily to introduce an out-of-phase component of control transformer rotor voltage which becomes quite dominant as the rotation speed approaches synchronous speed. At the lower speeds the major effect is to shift the zero voltage position, thus causing the error in transmission of about one degree per 320 rpm noted previously.

It is of interest to note that the effect of rotation on error is not a function of the absolute speed but rather the relative speed of rotation ($v\omega$) compared to the synchronous speed of rotation (ω) of the excitation frequency. Thus by doubling the excitation frequency, an error of one degree would be obtained at about 640 rpm instead of 320.

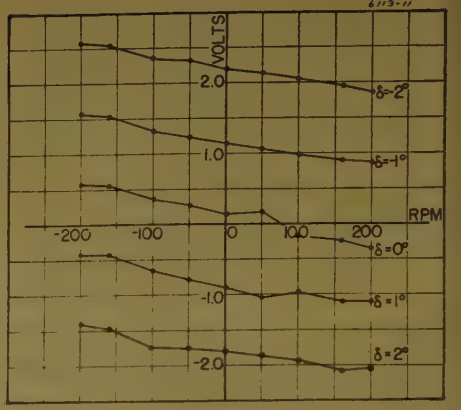


Figure 11. Averaged measured values of control transformer voltage versus speed

From test data. For generator-control transformer with fixed angular separation (δ)

Summary

The Selsyn generator-control transformer system has a high accuracy of the order of 0.3 degree. The static errors which are present because of resistance or reactance unbalance are evidenced by a second harmonic periodicity with rotor rotation. Errors caused by improper shaping of the rotor appear as higher harmonics of rotor position such as the sixth.

Even though a generator-control transformer system has no static errors, speed voltage effects produce a dynamic error which is of the order of one degree per 320 rpm for present-day 60-cycle machines.

Nomenclature

r_{ag} = line-to-line resistance in ohms of generator from terminal 1 to terminal 2

r_{bg} = line-to-line resistance in ohms of generator from terminal 1 to terminal 3

r_{abg} = resistance in ohms of phase 1 of generator

r_{ac} = line-to-line resistance in ohms of control transformer from terminal 1 to terminal 2

r_{bc} = line-to-line resistance in ohms of control transformer from terminal 1 to terminal 3

r_{abc} = resistance in ohms of phase 1 of control transformer

R_D = resistance in ohms of generator rotor

$r_{dg} = r_{qg}$ = per phase resistance in ohms of generator

$r_{dc} = r_{qc}$ = per phase resistance in ohms of control transformer

X_{ag} = line-to-line reactance in ohms of generator as seen by loop current i_a flowing from phase 2 to phase 1 with i_b not flowing. This reactance is a

second harmonic function of generator rotor position. Its numerical value may be obtained from a measurement of maximum and minimum values of the line-to-line reactance as viewed from terminals 1 and 2, with terminal 3 not connected. Expressed in terms of the per phase values of direct and quadrature axis reactances,

$X_{ag}=X_{dg}+X_{qg}-(X_{dg}-X_{qg})\times\cos(2\theta-120)$

X_{bg} = line-to-line reactance in ohms of generator as seen by loop current i_b flowing from phase 3 to phase 1. This reactance is a second harmonic function of generator rotor position, and its value may be obtained from a measurement of maximum and minimum values of line-to-line reactance as viewed from terminals 1 and 3.

$X_{bg}=X_{dg}+X_{qg}-(X_{dg}-X_{qg})\times\cos(2\theta+120)$

X_{abg} = mutual reactance in ohms of generator between circuit of loop current i_a and that of loop current i_b. Expressed in terms of the values of direct and quadrature reactances noted above,

$$X_{abg}=\frac{X_{dg}+X_{qg}}{2}-\frac{(X_{dg}-X_{qg})}{2}\times\cos[2\theta-120]$$

X_{ac} = line-to-line reactance in ohms of control transformer as seen by loop current i_a flowing from phase 1 to phase 2 with i_b not flowing. Because of the absence of reactance variation with a change in control transformer rotor position,

$X_{ac}=X_{dc}+X_{qc}$

X_{bc} = line-to-line reactance in ohms of control transformer as seen by loop current i_b flowing from phase 1 to phase 3 with i_a not flowing

$X_{bc}=X_{ac}=X_{dc}+X_{qc}$

X_{abc} = mutual reactance in ohms of control transformer between circuit of loop current i_a and that of loop current i_b

$$X_{abc}=\frac{X_{dc}+X_{qc}}{2}$$

X_{Dg} = reactance in ohms of generator rotor

X_{MDag} = mutual reactance between generator stator windings, phases 1 and 2 and the generator rotor winding. This reactance is a function of the odd harmonics of rotor position; however, because of the 3-phase arrangement of the stator windings, those harmonics which are multiples of three are missing. Expressed in terms of the maximum values of the fundamental, X_{M1g}, fifth, X_{M5g}, and seventh, X_{M7g}, harmonics of generator mutual reactances between the rotor and the line-to-line stator windings,

$X_{MDag}=X_{M1g}\cos(\theta+30)+X_{M5g}\times\cos(5\theta-30)+X_{M7g}\cos(7\theta+30)$

X_{MDbg} = mutual reactance between generator stator windings phases 1 and 3 and the generator rotor windings. It is an odd harmonic function of generator rotor position as X_{MDag}. Expressed in terms of X_{M1g}, X_{M5g}, and X_{M7g},

$X_{MDbg}=X_{M1g}\cos(\theta-30)+X_{M5g}\times\cos(5\theta+30)+X_{M7g}\cos(7\theta-30)$

X_{M1g} = maximum value in ohms of fundamental of mutual reactance between generator rotor and stator windings, line-to-line

X_{M5g} = maximum value in ohms of fifth harmonic of mutual reactance between generator rotor and stator windings, line-to-line

X_{M7g} = maximum value in ohms of seventh harmonic of mutual reactance between generator rotor and stator windings, line-to-line

X_{M1c} = maximum value of mutual reactance in ohms of control transformer between line-to-line of stator and rotor winding

X_{dg} = per phase value of direct axis reactance in ohms of generator stator windings. In terms of the maximum value of X_{ag},

$$X_{dg}=\frac{X_{ag}}{2}\text{ (maximum)}$$

X'_{dg} = symbol used to represent the apparent reactance of the stator direct axis as modified by the presence of the current flowing in the rotor

X_{qg} = per phase value of quadrature axis reactance in ohms of generator stator windings. In terms of the minimum value of X_{ag},

$$X_{qg}=\frac{X_{ag}}{2}\text{ (minimum)}$$

X_{dc} = per phase value of direct axis reactance in ohms of control-transformer stator windings. In terms of X_{ac},

$$X_{dc}=\frac{X_{ac}}{2}$$

X_{qc} = per phase value of quadrature axis reactance in ohms of control-transformer stator windings. In terms of X_{ac},

$$X_{qc}=\frac{X_{ac}}{2}$$

$\Delta(X_{ag}+X_{ac})$ = ohms value of unbalance in

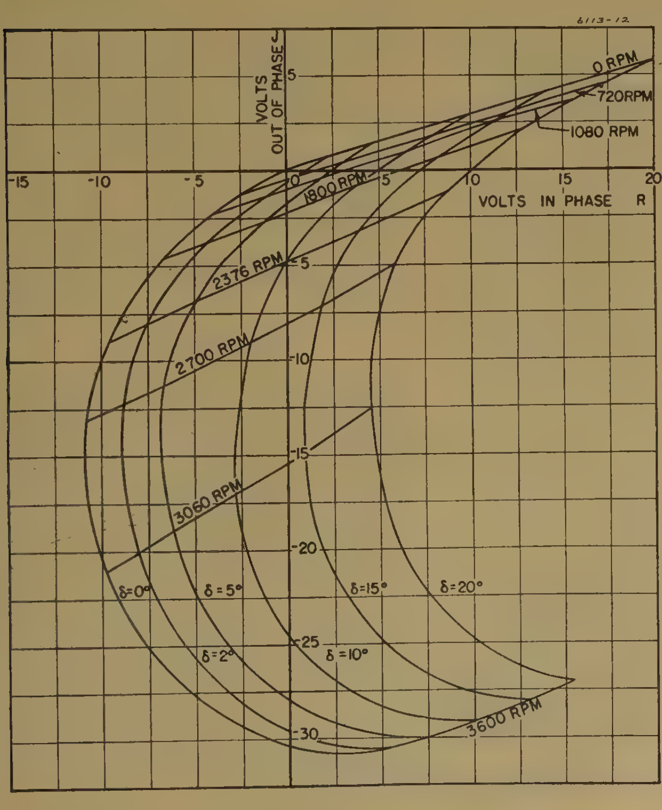
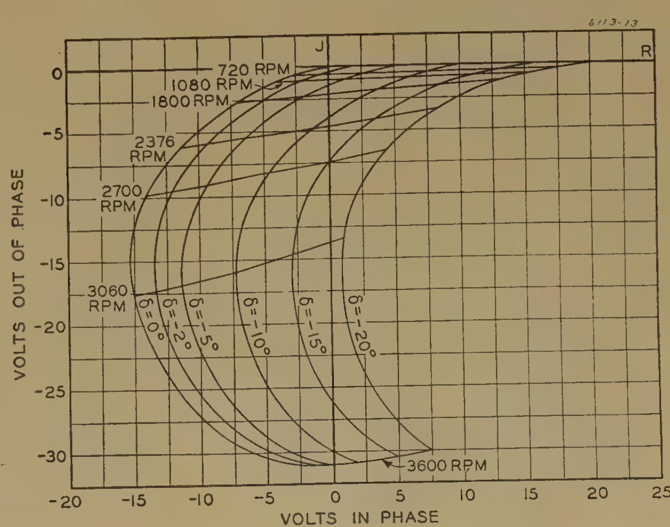


Figure 12 (left). Control transformer voltage components for constant velocity operation with fixed separation between rotors

Generator excitation voltage used as in-phase reference (R)

Figure 13 (below). Control transformer voltage components for constant velocity operation with fixed separation between rotors

Control transformer voltage at standstill is in-phase reference



loop reactance used for specifying error voltage ϵp as a function of reactance unbalance

i_a, i_b = loop currents in amperes flowing in generator-control transformer stator circuits as shown in Figure 3

I_D = generator rotor current in amperes flowing as shown in Figure 3

I_a = rms value of the magnitude of the loop current i_a

I_b = rms value of the magnitude of the loop current i_b

i_{dg} = direct axis current in amperes flowing in generator stator circuit shown in Figure 4

i_{qg} = quadrature axis current in amperes flowing in generator stator circuit shown in Figure 4

i_{dc} = direct axis current in amperes flowing in control-transformer stator circuit shown in Figure 4

i_{qc} = quadrature axis current in amperes flowing in control-transformer stator circuit shown in Figure 4

ψ_{dg} = direct axis flux linkages in the generator stator

ψ_{qg} = quadrature axis flux linkages in the generator stator

ψ_{Dg} = direct axis flux linkages in the generator rotor

E_g = rms value of the magnitude of the excitation voltage impressed on the generator rotor

E_c = rms value of the voltage present at the rotor of the control transformer. This is the so-called control transformer signal voltage

E_D = excitation voltage applied to generator rotor

E'_D = symbol used to represent the voltage present in the stator of the generator produced by the actual excitation voltage in the generator rotor

e_{dg} = per phase value of direct axis voltage in volts for generator stator

e_{qg} = per phase value of quadrature axis voltage in volts for generator stator

e_{dc} = per phase value of direct axis voltage in volts for control transformer stator

e_{qc} = per phase value of quadrature axis voltage in volts for control transformer stator

ϵp = position error in degrees between the control-transformer rotor position for zero volts output and the position 90 degrees displaced in a lagging direction from the generator rotor. See equation 5 for sign convention employed

ϵu = rms value of the magnitude of the out-of-phase component of control transformer rotor voltage for that rotor position where ϵp is zero

θ = angular position of generator rotor. In general, the units associated with θ are degrees, but in those cases where the term $p\theta$ is used, the units of θ are radians, so $p\theta$ equals an angular velocity in radians per second

α = angular position of control-transformer rotor. The same comments in regard to units apply to α as to θ

β = phase angle in degrees between the loop currents i_a and i_b. See equations (7) and (8) for the sign convention employed

ω = angular frequency in radians per second of generator excitation voltage. In the calculations, ω = 377 radians per second corresponding to 60 cycles per second

V = speed of generator and control transformer rotors expressed as a ratio to synchronous speed at excitation frequency ω

$\delta = \alpha - \theta$ = difference in angular position between generator and control

C, D, M, N = symbols used to replace the more complicated reactance expressions of equations 34–37. These are of use only in literal work and need not appear in the numerical calculations

t = time in seconds

$p = \frac{d}{dt}$ = derivative with respect to time

j = operator indicating $\sqrt{-1}$

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High Voltage D-C Testing of Rubber-Insulated Wire

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At different stages in the manufacture of insulated wire and cable, various electrical and physical tests are made to ensure the quality and serviceability of the completed cable. One of the most important of these tests is the high voltage test, a-c or d-c. It may be applied to short sections of the insulated conductor as it passes through electrodes of various types, dry or wet, or to full factory lengths of the insulated wire, dry or wet.

The scope of this paper is limited to the results of an experimental study made on the voltage limits suitable for the high voltage d-c testing of factory lengths of plain insulated wire immersed in water. In addition to many sample tests, the study included tests on more than 5,500 different factory lengths of submarine cable conductor, insulated with natural rubber, synthetic rubber, or polyethylene. Such factors as testing voltage, type of fault, and immersion time were varied over a wide range. From these data test voltage limits are suggested; the lowest test voltage that will indicate serious defects and the highest test voltage that can be used without causing damage to sound insulation, from either the d-c test voltage itself or from impulse voltages incidental to the testing.

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In the following discussion all insulating compounds are of the submarine type, all wire sizes are American Wire Gauge and all voltages are given as volts per mil thickness of insulation (total voltage per mils thickness). All alternating voltages are root-mean-square. Unless otherwise indicated, all d-c tests were made with negative polarity on the conductor.

Dielectric Strength of Rubber and Polyethylene Insulated Wire

Knowledge of the d-c and impulse dielectric strength is essential to the proper selection of a suitable d-c test voltage. In Table I are given typical values for the dielectric strength of the materials covered in this discussion. They represent many tests made over several years time. All test specimens were five feet active length tested after 24 hours immersion in room temperature water. The

calculated impulse wave was 1.5x40 microseconds. The d-c tests were made with a voltage increase of 700 volts per second. The alternating voltage was increased 500 volts per second. The wire size varied over a moderate range in the neighborhood of number 12 6/64 inch thickness.

Damage to Sound Insulation by Excessive Test Voltage

The tests reported in Tables II and III were made to investigate the possibility that the application of high d-c test voltage might damage sound insulation by causing a permanent decrease in the dielectric strength.

The tests in Table II were made on insulated wire samples of five feet active length after 24 to 48 hours immersion in room temperature water. All samples of the same size were cut from the same length and grouped in such a way as to minimize the effect of variation of dielectric strength along the length. After a 30-minute application of the direct voltages indicated in the table the samples were tested for a-c dielectric strength with a voltage increase of 500 volts per second. In all cases the a-c tests were made the day after the d-c tests, without removing the samples from the water.

The tests in Table III were made on number 12 solid, 6/64 inch thickness of natural rubber. A 175-foot length in water was tested at 640 volts d-c per mil for five minutes without failure and then cut into samples of 5-foot active length

Table I. Dielectric Strength of Rubber and Polyethylene Insulated Wire

Type of Insulation	Impulse Strength in Volts Per Mil	D-C Strength in Volts Per Mil	A-C Strength in Volts Per Mil
Natural rubber	1,070	1,500	350
GR-S rubber	1,030	1,400	370
Polyethylene	1,500	3,000	350

for the tests indicated in the table. The dielectric strength tests were made with a direct voltage increase of 700 volts per second and an impulse wave of 1.5x40 microseconds. There were five samples in each of the four groups. In other respects the testing conditions were similar to those of Table II.

The results in Tables II and III show no real decrease in a-c or d-c strength resulting from previous applications of direct or impulse voltage. None of the variations apparent in the tables exceeds those usually found in such a collection

Table II. Effect of High Voltage D-C Tests on Subsequent A-C Strength

Size	Previous Test	Number of Samples	60-Cycle Breakdown Voltage: Highest Volts Per Mil	Lowest Volts Per Mil	Average Volts Per Mil	Average in Per Cent of No D-C Test
Number 12 (7 strands)—6/64 inch thickness polyethylene	None	15	361	278	326	100
	800 volts per mil d-c	5	350	317	338	104
	1,600 volts per mil d-c	10	333	266	301	93
Number 6 (7 strands)—10/64 inch thickness GR-S rubber	None	15	350	293	319	100
	500 volts per mil d-c	5	363	286	313	98
	800 volts per mil d-c	10	388	300	334	105
Number 12 solid—6/64 inch thickness natural rubber	None	13	455	344	411	100
	500 volts per mil d-c	5	419	328	374	91
	800 volts per mil d-c	8	415	338	378	92
Number 6 (7 strands)—8/64 inch thickness natural rubber	None	5	445	361	422	100
	800 volts per mil d-c	5	470	378	440	104

of dielectric strength data. Therefore, the use of d-c testing voltages as high as one half of the 5-foot d-c dielectric strength in Table I should not result in any appreciable decrease in the dielectric strength of sound insulation.

Relation Between Type of Fault and Necessary Test Voltage

The object of high voltage testing is to indicate weak spots or partial faults in the insulation by applying enough voltage to cause the complete failure of such spots. Therefore, it is desirable to know something of the relation between the type of the partial faults and the test voltage necessary to develop such spots into complete failure.

In Table IV are given the results of voltage tests made on artificial faults in number 12 standard, 6/64 inch thickness natural rubber. After tying the wire to a wooden splint each fault was made by forcing a needle or revolving drill through the insulation to the conductor as indicated by completion of a buzzer circuit, then withdrawing the needle or drill. These faults were tested with high direct voltage after immersion in water for three days. The results in Table IV indicate that such faults may withstand direct voltage as high as 225 volts per mil of insulation for 40 minutes (20 minutes each polarity) without failure. Thus, a test voltage as low as this may not indicate faults that extend entirely through the insulation thickness. Open artificial faults were made in the same size wire by cutting away part of the insulation thickness at one spot. High voltage d-c tests on these faults indicated that a remaining thickness of 3/64 inch, or only one half the total thickness, is enough to withstand 500 volts d-c per mil of the total thickness without failure. A similar relation also is indicated by comparison of the test voltage with the dielectric strength values in Table I. For instance, a test voltage of 500 volts d-c per mil is only one third the d-c dielectric strength of natural rubber. Thus, a small part of the insulation at a single spot might withstand a test voltage of 500 volts per mil total thickness.

When rubber insulated wire is wound on a reel there may be considerable pressure between adjacent turns of wire. The results in Table V show the effect of such pressure on the d-c dielectric strength of artificial faults in wire, number 12, 6/64 inch thickness natural rubber. The d-c strength of needle faults in water was increased from 150 to 690 volts per mil, by applying pressure from adjacent rubber insulated wire as on a reel.

Also in Table V are the results of similar tests on a fault in the same size wire consisting of a *U*-shaped knife cut all the way to the conductor. After immersion

Table III. Effect of High Voltage D-C and Impulse Tests on Subsequent D-C Strength

Previous Test	D-C Dielectric Strength Volts Per Mil: Highest	Lowest	Average
a—None	1,450	1,290	1,360
b—640 volts per mil d-c for 30 minutes	1,500	1,120	1,340
c—640 volts per mil d-c for 5 minutes, 6 times	1,480	1,310	1,360
d—850 volts per mil impulse, 6 times	1,530	1,360	1,470

in water for 24 hours under pressure from adjacent rubber insulated wire this fault had a d-c strength of 600 volts per mil and after the first failure a voltage of 480 volts per mil was necessary to cause a second failure.

In addition to these data on artificial faults two examples of natural faults are offered:

1. Three thousand feet of number 14 solid 3/64 inch thickness of natural rubber withstood, without failure, three voltage tests

five minutes each at 340 volts d-c per mil during a 12-day water immersion. After six months' additional room temperature immersion the wire had an insulation resistance of less than 10 megohms per 1,000 feet at 500 volts d-c because of two air pockets in the insulation so large that only ten per cent of the wall remained in a sound condition.

2. Ten thousand feet of number 12 standard, 6/64 inch thickness GR-S rubber withstood, without failure, three voltage tests five minutes each at 570 volts d-c per mil during a 12-day immersion. Subsequent immersion and high voltage testing finally disclosed an iron chip so located within the insulation that only one third of the thickness was left effective.

These data on both artificial and natural faults indicate that under some circumstances it is possible for serious faults to withstand several applications of d-c testing voltage as high as 500 volts per mil without failure.

Damage to Sound Insulation by Impulse Voltages Incidental to High Voltage D-C Testing

When factory lengths of insulated wire are at the d-c test voltage, abrupt grounding of the conductor is likely to produce momentary voltages that are higher than the test voltage. If these voltages exceed the impulse strength of the insulation, failures will ensue. There are three principal ways in which the conductor may be grounded abruptly:

1. Accidental flashover during test.
2. Intentional grounding after completion of test.
3. Failure of insulation on test.

The first way can be prevented by care and the second by the use of series resistance but limitation of the test voltage is the only known method of preventing damage from the third cause.

In a total of more than 17,000 voltage tests on more than 5,500 factory lengths with an insulation thickness in the general neighborhood of 6/64 inch and with different test voltages between 300 and 620 volts per mil on rubber and between 500 and 1,000 volts per mil on polyethylene, impulse failures have been limited to test voltages of 580 volts per mil and higher on natural or synthetic rubber and to 850 volts per mil and higher on polyethylene.

Investigation of these impulse voltages with cathode-ray oscillograph at direct voltages less than 100 volts indicates that such voltages may get as high as 1.8 of the applied voltage (test voltage). Application of this ratio to the impulse strengths in Table I gives minimum test voltages of 593, 573, and 835 volt per mil for natural rubber, GR-S rubber, and polyethylene respectively, at which impulse failures might be expected. These indications admittedly are not conclusive because of the low test voltage but it is interesting to note the close agreement with the testing experience reported in the preceding paragraph.

Table IV. Dielectric Strength of Artificial Faults

	D-C Strength in Volts Per Mil Thickness					
Needle diameter, inch	0.025	0.025	0.025	0.050	0.057	0.042
Type of needle point	Extra fine	Fine	Medium	Medium	Blunt	*
	205	159	142	200	68	27
	225†	230	159			
	285**†					

* Hole made with 0.042-inch twist drill.

** Insulation stretched by bending wire before puncturing insulation. Wire then straightened and bound to a splint.

† No failure after 40 minutes on 20 kv (5 minutes plus, 5 minutes minus, 15 minutes plus, 15 minutes minus).

Conclusions

High voltage d-c testing has been found of considerable value in controlling the quality of insulated wire but no implication is intended in this paper that such testing is essential to the maintenance of satisfactory quality. Comparison of high voltage d-c testing of factory lengths with other kinds of testing is outside the paper's limited scope. If high voltage d-c tests are made on factory lengths it would be well to consider the following voltage limitations that have been indicated by experimental study to date.

If the d-c test voltage used on factory lengths is too high there is a possibility of developing enough impulse voltage to cause failures in sound insulation.

As a value for the highest d-c test voltage at which there is no danger from impulses one of the following is suggested:

1. Fifty per cent of the 5-foot impulse strength.
2. Five hundred volts per mil on rubber and 750 volts per mil on polyethylene if the impulse strength is not known.

Prolonged or repeated applications of d-c test voltage as high as 50 per cent of the 5-foot d-c strength do not cause any reduction in the dielectric strength of sound insulation. This value is above the maximum limitation in test voltage imposed by the impulse strength of the insulation.

The dielectric strength of some faults is surprisingly high and may be increased further by pressure from adjacent wire. For this reason 25 per cent of the 5-foot d-c strength is suggested as the lowest voltage at which such tests should be made. The faults described and tested in the foregoing are not common in well-organized factories but it is possible for such faults to withstand tests as high as 33 per cent of the 5-foot d-c strength without failure.

Table V. Effect of Pressure on Strength of Artificial Faults

Test Specimen	D-C Strength in Volts Per Mil After 24 Hours in Water
Needle fault with no pressure	135
Needle fault squeezed against one adjacent wire without fault by wooden clamp	170 / 310
Needle fault squeezed in group with three other wires without faults, by binding with cotton string	340 / 690
U-knife cut fault squeezed in group with 6 other wires without foregoing faults	600 then 480

Because of these voltage limitations to high voltage d-c testing it is important to increase the effectiveness of the testing as much as possible, by such means as increase of the soaking time and decrease of the pressure from adjacent wire.

Dangerous Electric Currents

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Synopsis: This paper discusses lethal electric currents and their accompanying physiological effects, and interprets the data of a previous paper[4] in accordance with an original method of analysis found useful by the author in his own investigations of let-go currents. The present analysis concerns itself with threshold currents likely to produce instantaneous electrocution in one-half per cent of a large group of normal men. Although the conclusions are derived from tests made on animals, it is believed that the results may be indicative of what might be expected to occur in man. The majority of the work is based on experiments made at 60 cycles with shock durations of 0.03 to 3.0 seconds. Predictions of lethal currents for both direct current and capacitor discharges, while more speculative because of the limited data available, are included because of their importance due to the greatly increased use of direct current and electronic equipment.

PROGRESS in the development of electric equipment has brought increasing demands for additional knowledge on the effects of electric shock, particularly with regard to the maximum current that man reasonably might be expected to withstand without fatal results. Sensations, muscular contractions, and the current required to cause the victim to freeze to a circuit have been covered fairly well in recent papers on let-go currents.[1,2,3] The object of this paper is to extend the analyses and discuss effects at higher currents.

Any definition of lethal currents must consider the following factors:

1. Current pathway through the body.
2. Physical condition of the victim.
3. Magnitude of the current.
4. Shock duration.
5. Frequency.
6. Wave form.
7. The phase of the heart cycle at the instant the shock occurs.

Electric shocks produce different effects depending upon the structure through which the current passes. Currents flowing in the region of the heart may produce a condition of the heart muscle known as ventricular fibrillation. This condition usually is fatal and commonly is referred to as instantaneous electrocution. Currents passing through nerve centers controlling breathing may produce respiratory inhibition, that is, stoppage of respiration. Failure of the breathing mechanism often is temporary, the paralysis lasting from a few minutes to a few hours after interruption of the current. These victims often may be saved by prompt application of artificial respiration. Currents passed across the head from temple to temple may produce unconsciousness and muscular convulsions. Shocks of this type are used currently in electric shock treatment in certain types of insanity. If the current pathway involves lower nerve centers, the shock might produce ejaculation. Electric ejaculation and artificial insemination apparently have promise in breeding animals. If the pathway involves only an extremity, such as a hand or leg, memory of a disagreeable experience might be the only lasting result. Because of such variations in the effects of electric shock, it is customary to discuss lethal currents with respect to the most dangerous current pathway likely to be experienced in accidents. This is a pathway involving the heart, with external contacts usually assumed between the hands, or between one hand and the opposite foot.

Much attention has been given the study of fatal accidents. These studies have been very helpful in developing practical safety procedures and safety codes. From a technical point of view, results have been qualitative rather than quantitative. This is largely because of the difficulty of determining accurately the many variables involved. In addition to the factors already mentioned, it usually is possible only to make rough estimates in arriving at values of circuit impedance, body and contact resistances, and elapsed time between occurrence of the accident, rescue, and resuscitation.

Often these uncertainties have resulted in erroneous conclusions and have confused the issue. Perhaps the most serious misconception concerns the effects of voltage versus the effects of current. Current and *not* voltage is the proper criterion of shock intensity. The remainder of this paper will be devoted to a discussion of the electric shock hazards, caused by currents of various shock durations, wave shapes, and frequencies.

Sixty-cycle currents at the let-go level, that is, currents of from 10 to 20 milliamperes from hand to hand, are painful and hard to endure for even a short time. If long continued, they may lead to collapse, unconsciousness, and death. The physical condition of the victim is of prime importance in determining this hazard. The muscular contractions and accompanying sensations increase in severity as the current is increased. The muscular contractions progress up the arm to the chest until they become so severe that the victim is unable to breathe. Obviously, if the current flows for more than a few minutes, death may result from asphyxiation. However, if the circuit is interrupted in a reasonable time, breathing resumes automatically, and no serious after-effects result. Currents considerably in excess of those required to cause a stoppage of breathing by excessive muscular contraction of the chest muscles may produce a temporary paralysis of respiration by action on the nerves. It has been known for some time that respiration might be inhibited by currents passing through the respiratory nerve center located in the base of the brain. Observation of temporary muscular paralysis in human accidents has prompted Howard Miller, Southern California Edison Company, to suggest that respiratory inhibition also might be caused by currents affecting the nerve centers controlling the diaphragm. The suggestion is in agreement with W. Einthoven, who some 30 years ago demonstrated that electric currents applied directly to a nerve, insufficient to cause permanent damage, could produce complete blocking of the nerve for periods of the order of one-half hour. The respiratory paralysis may last for a considerable period after interruption of the current. In this case, resuscitation must be applied immediately to prevent asphyxial death. Often the paralysis disappears in a few minutes or in a few hours, and continued application of artificial respiration may save the victim. Mere cessation of natural breathing is not likely to produce serious aftereffects or permanent damage, as evidenced by the many persons who have been resuscitated successfully.

Much valuable information on currents producing ventricular fibrillation

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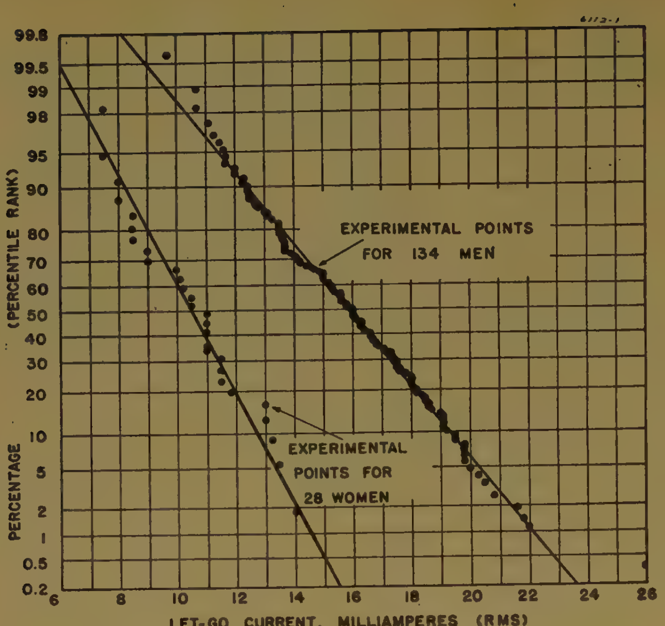


Figure 1. Sixty-cycle sine wave let-go current distribution curves for men and women

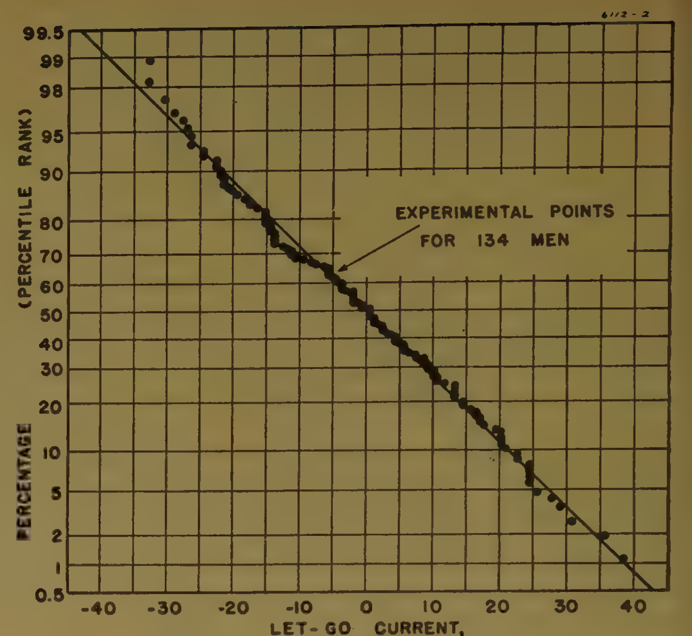


Figure 2. Sixty-cycle sine wave let-go current deviation curve for men

has been obtained from extensive experiments performed on animals at Columbia University by Ferris, King, Spence, and Williams.[4] These authorities adeptly have described the fibrillating condition as follows:

"Currents somewhat greater than those just necessary to stop respiration by action on the muscles may cause fatalities, even though the duration of such shocks is but a few seconds or less—far too short to be important from the standpoint of interruption of respiration, and obviously too short to give any opportunity for rescue before the end of the shock. Death under such conditions is brought about by ventricular fibrillation, which is a disruption of normal heart action. This condition is an uncoordinated asynchronous contraction of the ventricular muscle fibers in contrast to their normal co-ordinated and rhythmic contraction. It results from an abnormal stimulation rather than from damage to the heart. In the fibrillating condition, the heart seems to quiver rather than to beat; no heart-sounds can be heard with a stethoscope—the pumping action of the heart ceases; failure of circulation results in asphyxial death within a few minutes."

Most of the experiments were made on sheep, calves, pigs, and dogs, in which chest dimensions, body weights, heart weights, and heart rates were comparable to those of man, although several species of smaller animals, including guinea pigs, rabbits, and cats were included to establish the general trend of effects with weight and other physiological factors. Foremost of their findings from the practical point of view are:

1. The susceptibility of the heart to fibrillate depends on the phase of the heart cycle at the instant the shock is applied.

2. Repeated shocks are not cumulative in their effects on the heart, and the heart generally returns to normal in about five minutes if fibrillation does not occur.

3. The results on the whole indicated that sinusoidal currents in excess of 100 milliamperes at 60 cycles from hand to foot would be dangerous for shock durations of three seconds or more.

The statement establishing the threshold current producing ventricular fibrillation in man at 100 milliamperes for 60-cycle sine-wave shocks of three seconds or more has been accepted quite generally by the profession. Without any thought of depreciating the value of this conclusion, it is fair to say that much is to be desired yet in the way of additional information, with special regard to the variation of effects with shock duration, wave form, and frequency.

Before proceeding with the proposed analysis of fibrillating currents, it is helpful to review briefly certain conclusions found in the let-go current investigations.[2,3]

Experimental points representing let-go currents for 134 men and 28 women are shown in Figure 1, in which the ordinate gives the percentage of subjects who can release their grasp of a conductor carrying the current shown in the abscissa.[2,3] It is important to emphasize that a sufficient number of points was obtained to determine a normal distribution, as evidenced by the fact that the data closely follow a straight line when plotted on probability paper. Figure 2 shows the results for the 134 men plotted as per cent deviations from 15.87 milliamperes, the mean of the group. The straight line governed by the majority of the points of Figure 2 has been called the deviation curve. When points representing the women were plotted as per cent deviation from the mean of their group, it was found that they followed a deviation curve having the same slope. Similar results were found for composite waves consisting of a mixture of alternating current and direct current, for rectified 60-cycle sine waves, and for sine-wave alternating currents from 5 to 5,000 cycles. This finding is important because it permits improved accuracy in predicting results for large groups based on a relatively small number of experimental points.

This procedure will be used now to study 60-cycle fibrillating currents obtained by Ferris and his associates. Figure 3 shows fibrillating currents plotted as per cent deviations from the mean of each series, respectively, for eight different tests. Although the distribution of the points does not line up as well as that found for let-go currents, it does establish that a definite trend exists. This holds for the four species of animals for 3-second shocks, and for the sheep for shock durations of 0.03 to 3.0 seconds. It is possible that the same trend also might hold for shocks of various durations for the other animals and for man.

Because of the small number of points available for any test, the straight line governed by all the points was drawn to determine the deviation curve. Apparent discrepancies of a few of the points

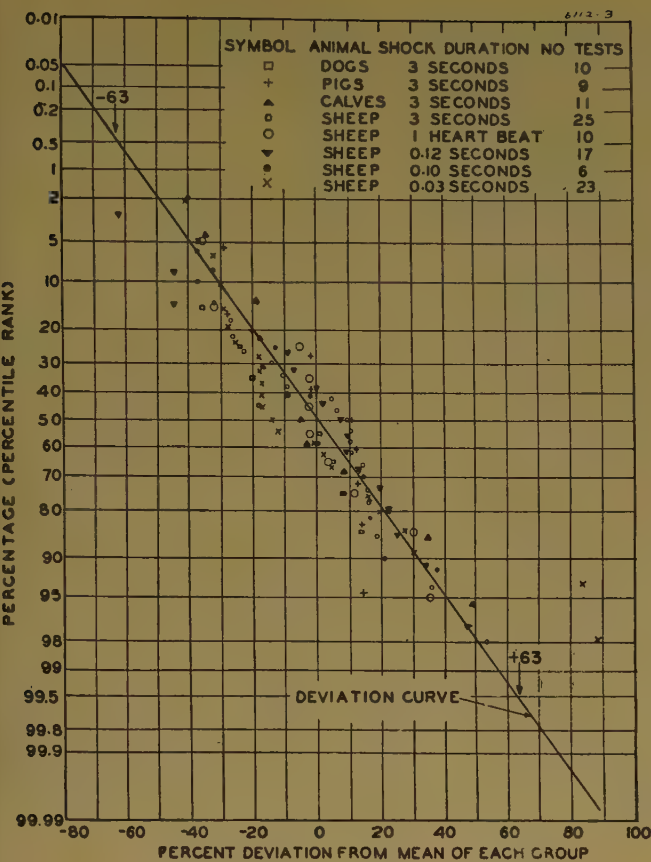


Figure 3. Sixty-cycle sine wave fibrillating current deviation curve for various animals

from the deviation curve are believed to be caused by the small number available for any single test, to experimental errors in applying the shocks at the most susceptible phase of the heart cycle, to the rather wide variation in body weights, and to other factors which properly are included in the term biological variability.

The fibrillating current for an animal was taken as 1/2 (minimum current causing fibrillation + maximum current not causing fibrillation). It is believed that the value so obtained should be reasonably close to the current just required to produce fibrillation during the most susceptible phase of the heart cycle, and the statistical method of analysis should minimize errors.

The current pathway for these experiments was between the right foreleg and the left hind leg. Experiments also were made using four other pathways, namely, across the chest, chest to foreleg, head to hind leg, and between both hind legs. For the current pathway between the two hind legs, the portion of current reaching the heart was evidently too small to produce fibrillation for currents within the operating range of the equipment. Differences in values for the other pathways did not appear great enough to be significant, and it was concluded that results obtained for the pathway between the foreleg and the hind leg should be sufficiently accurate for most engineering purposes.

Data obtained by the investigators at Columbia University[4] for 129 points were furnished the author. Of this number, 111 were used and 18 omitted in the statistical analysis. The points rejected represented two different conditions. Eight high points, designated by these investigators as mode *A*, were omitted to give conservative results. The remaining ten points were from the 0.47-second shock test made on sheep.

The following is submitted as justifying rejection of the 0.47-second test:

1. Results of the 0.47-second shock test are shown in Figure 4 in which the response follows a deviation curve of much smaller slope than that found for the other tests. Wide differences in the results of two almost identical tests suggest the possibility of error. The points representing shocks of one heartbeat (large open dots, Figure 3) are consistent with all the other tests, and fall very nicely about the deviation curve. The duration of the shocks for this test was adjusted to equal as nearly as possible the time of the individual's heartbeat and varied from 0.36 to 0.55 second, averaging 0.45 second. It is untenable that a difference of only 0.02 second in the average shock duration would be sufficient to cause the difference in response indicated by the two deviation curves.

2. The accuracy of analyses based on statistical procedures depends upon a representative number of points. It was difficult to establish deviation curves for let-go currents for less than about 25 points. The apparent inconsistency of a single test involving only 10 points therefore is not considered sufficient to invalidate the analysis.

Fibrillating currents for three-second shocks for seven species of animals versus body weight are shown in Figure 5. Points representing the larger animals total 55, and none were discarded. The smaller animals were represented by average values only, as data for these cases were not furnished the author. The 50 per cent line was drawn by inspection through the points representing the means of the various tests. The author is still in doubt as to where to draw the best line, and the analysis was carried on from here, using a broken line neglecting the points for the pigs, and a solid line in which the effect of the pigs was weighed by eye. The fact that the upper line passes very close to the points representing the averages for each of the other six tests is believed significant. It is unfortunate that only nine pigs were used.

In spite of the wide scattering of individual points, it is evident that in general the fibrillating current is proportional to body-weight. Similar variations were encountered when studying let-go currents, but all efforts to find a correlation with age, weight, strength of grip, or arm measurements, were without conclusive results; however, higher values were obtained on well-developed individuals having the appearance of good health. Other things being equal, it is believed that both let-go currents and

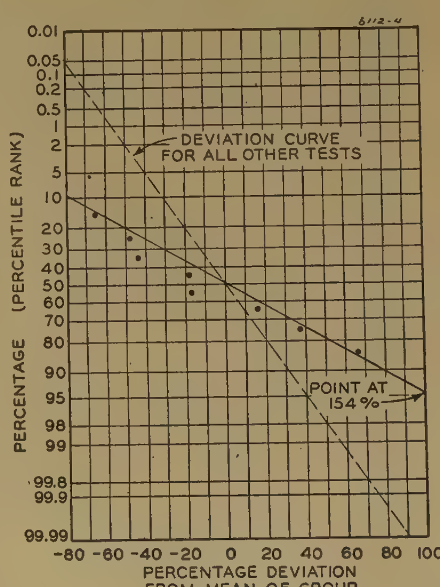


Figure 4. 60-Cycle sine wave fibrillating current deviation curve for sheep

Shock duration 0.47 second
Ten cases tested

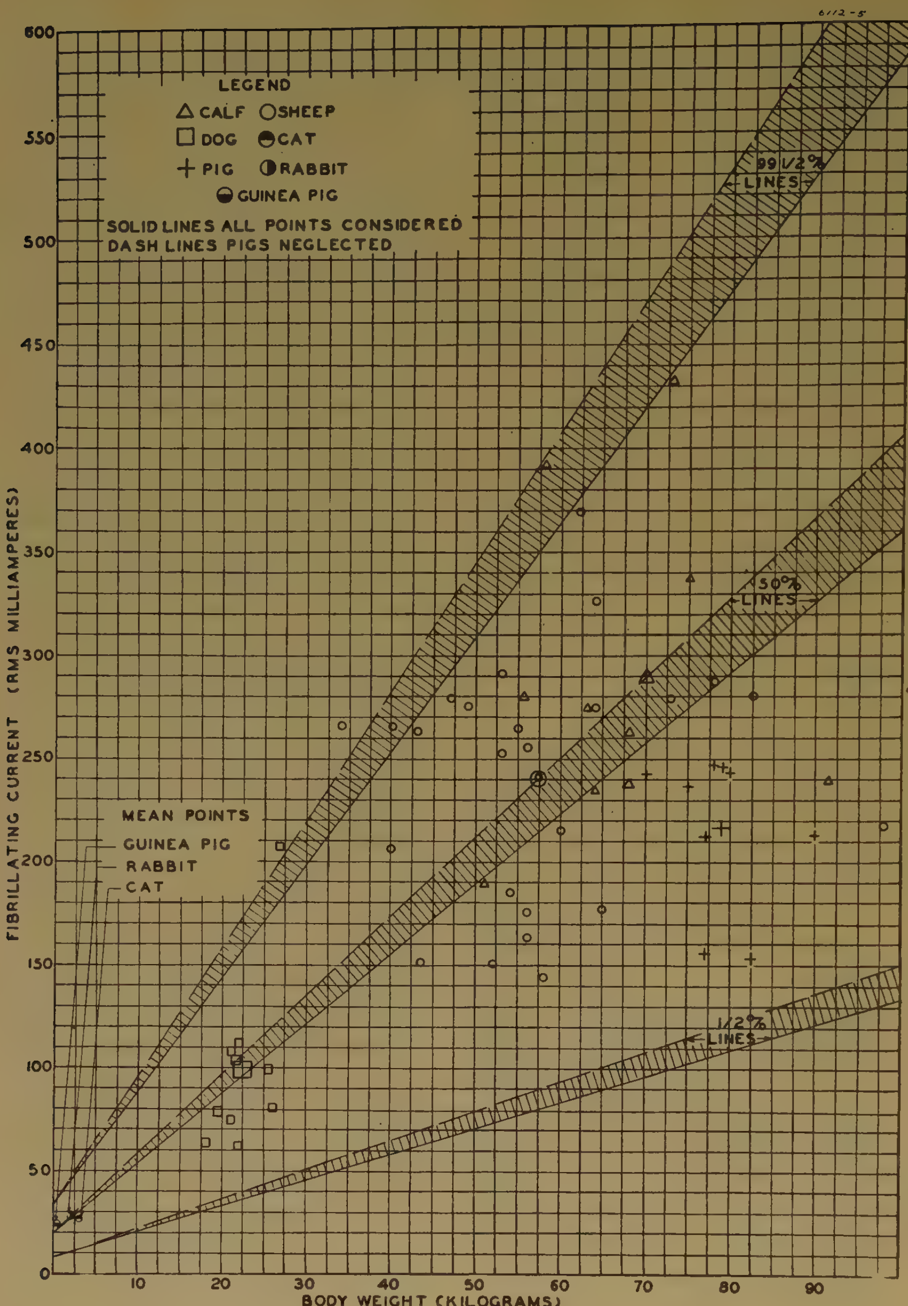


Figure 5. Relation of 60-cycle sine wave fibrillating current of various animals to body weight

Shock duration 3.0 seconds

fibrillating currents at least should be roughly proportional to size and weight.

The $99^1/_2$ and 1/2 per cent lines of Figure 5 were computed using equation 1, reference 2, and the deviation curve of Figure 3. The currents corresponding to a given percentile rank = mean of sample $\times$ (1 $\pm$ deviation from mean). The deviation from the mean for $99^1/_2$ per cent and for 1/2 per cent = ± 0.63, hence:

$$I(99^1/_2 \text{ per cent and } ^1/_2 \text{ per cent}) = I(50 \text{ per cent})(1 \pm 0.63) \quad (1)$$

Assume that the analysis applies to man. Enter the graph at 70 kilograms which is the weight of an average man, and proceeding vertically indicates that 60-cycle sine-wave alternating currents between 95 and 107 milliamperes and three seconds duration might produce ventricular fibrillation in one half of one per cent of a large group of normal men. This is in close agreement with the generally accepted figure of 100 milliamperes proposed by Ferris, King, Spence, and Williams.[4]

The equation for the lower one half per cent line may be written:

$$I(^1/_2 \text{ per cent}) = 1.26W + 7.4 \text{ milliamperes} \quad (2)$$

in which W denotes body weight in kilograms.

Figure 6 shows the results obtained on sheep plotted on log–log graph paper in the attempt to determine a relation between fibrillating current and shock duration. A total of 99 points is shown in this figure, including the rejected points, mode A and the 0.47-second shocks previously discussed. Because of the rather wide spread in the body-weight of the various animals, calculated points, shown as open dots on Figure 6, were corrected to 57.4 kilograms to obtain a standard reference. The 1/2 per cent and $99^1/_2$ per cent points were calculated using equations 1 and 6. The $99^1/_2$, 50, and 1/2 per cent lines were drawn with emphasis given the 0.03- and 3.0-second tests because of the greater number of animals used, and to obtain conservative results. Attention is directed to the lone point falling below the one half of one per cent line for the 0.47-second test. This exception must be given serious consideration in deciding whether or not the proposed analysis is acceptable.

If we assume a straight line may be used conservatively to represent the general trend, equations for the lines of Figure 6 may be expressed:

$$I = KT^n,$$

where

n = slope = $-^1/_2$
T = shock duration in seconds

Therefore

$$I = \frac{K}{\sqrt{T}} \qquad K = I\sqrt{T} \quad (3)$$

The equation for I(1/2 per cent) for 57.4-kilogram sheep is found as follows. From Figure 5, when T = 3 seconds, I(1/2 per cent) = 89.5 milliamperes. Substituting in equation 3,

$$\left.\begin{array}{l} K = 89.5\sqrt{3} = 155 \text{ milliamperes} \\ I(^1/_2 \text{ per cent}) \text{ sheep} \\ \quad = \dfrac{155}{\sqrt{T}} \text{ milliamperes} \end{array}\right\} \quad (4)$$

The I(1/2 per cent) equation for all 57.4-kilogram animals is found as follows. From Figure 5, I(50 per cent) sheep = 240.5 milliamperes, and W = 57.4 kilograms. The mean fibrillating current for the solid line for a weight of 57.4 kilograms = 215 milliamperes. The corresponding 1/2 percentile value is $89 \times 215/240.5 = 79.6$ milliamperes. Therefore,

$$K = 79.6\sqrt{3} = 138 \text{ from equation 3}$$

Therefore, for all 57.4-kilogram animals tested, including sheep,

$$I(^1/_2 \text{ per cent}) = \frac{138}{\sqrt{T}} \text{ milliamperes} \quad (5)$$

Equations for I(1/2 per cent), including effects of shock duration and various body-weights, are obtained as follows. Let subscripts 1 and 2 be used to represent animals 1 and 2, respectively. Equation 2 then may be written

$$\frac{I_1(^1/_2 \text{ per cent})}{I_2(^1/_2 \text{ per cent})}=\frac{1.26W_1+7.4}{1.26W_2+7.4}$$

or

$$I_1(^1/_2 \text{ per cent})=I_2(^1/_2 \text{ per cent})\times \frac{1.26W_1+7.4}{1.26W_2+7.4} \text{ milliamperes} \quad (6)$$

Substituting equation 5 in equation 6,

$$I(^1/_2 \text{ per cent})=\frac{138}{\sqrt{T}}\times \frac{1.26W+7.4}{1.26\times 57.4+7.4} \text{ milliamperes} = \frac{2.18W+12.8}{\sqrt{T}} \text{ milliamperes}$$

Inserting $W=70$ kilograms (average weight for man),

$$I(^1/_2 \text{ per cent}) \text{ fibrillating current for man} = \frac{165}{\sqrt{T}} \text{ milliamperes} \quad (7)$$

The broken line (neglecting the effect of the pigs) of Figure 6 was obtained by substituting equation 4 in equation 6 and $W=$ 70 kilograms. It is interesting to note that the uncertainty of the current versus weight-lines of Figure 5 produces an inappreciable difference in the final result, as indicated by the small width of the crosshatched area of Figure 6.

It should be repeated that this analysis is based on experiments in which the duration of the shock was varied from 0.03 to 3.0 seconds. Extrapolation for larger range of shock durations should be made with caution; possibly the relations might hold for durations from five seconds to one cycle or possibly one fourth cycle. No account is taken here of inhibition of respiration or of the cumulative effects of shocks at intervals of the order of seconds. These relations are based on the assumption that the heart recovers fully from one shock before a second shock occurs. In other words, the threshold currents are for single shocks. No data are available regarding the effects of repeated shocks such as those produced by intermittent electric fence controllers.[5,6] It must be stressed that application to man is entirely conjectural.

Relatively large currents (amperes and not milliamperes) may produce sufficient heat to destroy nerves, protoplasm, bone and cause hemorrhages resulting in immediate death. Delayed death may

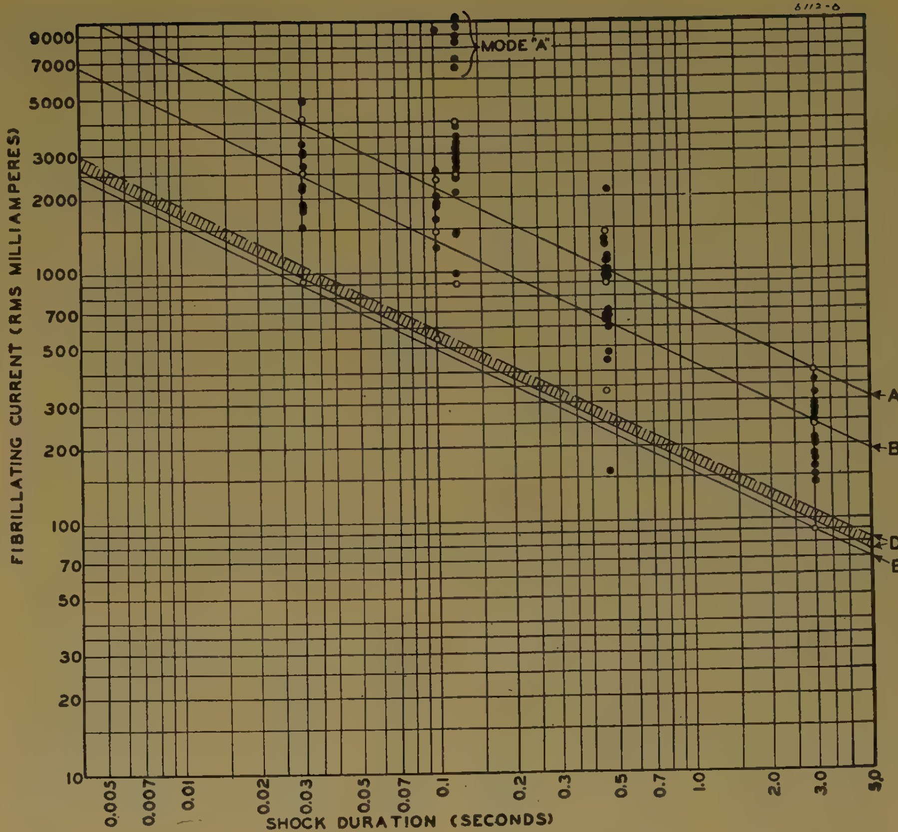


Figure 6. Relation of 60-cycle sine wave fibrillating current to shock duration

• Experimental points
o Calculated points
A—99½ per cent line for 57.4-kilogram sheep
B—50 per cent line for 57.4-kilogram sheep
C—½ per cent line for all 70-kilogram animals including man
D—½ per cent line for 57.4-kilogram sheep

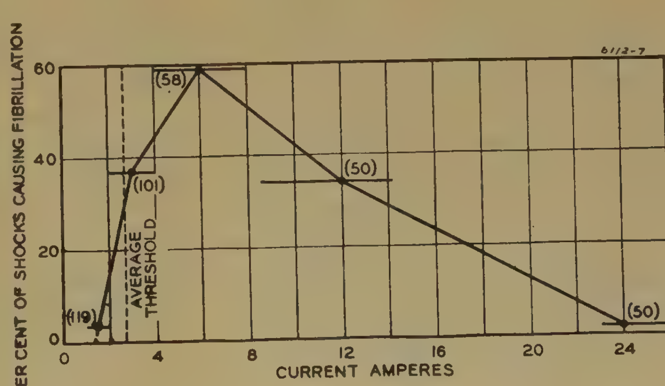
be due to burns or other complications.

For 0.03-second shocks, the hearts of sheep seemed to be most susceptible to fibrillate at currents of about six amperes, and less sensitive as the current was either increased or decreased from this value. Figure 7 was taken from reference 4 to illustrate the point. The explanation of this phenomenon is that high currents cause complete contraction of the entire heart musculature, and fibrillation is prevented. If the shock is of appreciable duration, death is inevitable. However, if the shock is of short duration, when the current is interrupted, relaxation may be followed by a spontaneous resumption of normal rhythmic contractions. It is believed that the abdominal massage and accompanying stimulation of the heart caused by the application of artificial respiration may be beneficial in assisting the heart to regain its normal rhythm. This is offered in explanation of infrequent accident cases in which victims apparently withstand relatively large currents.

Ventricular fibrillation may be caused by mechanical stimulation of the heart when it is exposed during surgical proce-

Figure 7. Effect of current on susceptibility of sheep hearts to ventricular fibrillation

Shock duration 0.03 second frequency 60 cycles. Shocks applied in partial refractory period of cardiac cycle. Number of shocks and current spread indicated for each point on curve



dures as well as by electric shock, and occasionally the heart goes into fibrillation during major operations involving the chest. Although several investigators have reported a degree of success in using counter-electric shocks to arrest fibrillation in animals, C. S. Beck[7] is the first to publish a procedure for human beings. He reports two cases in which two large electrodes were held directly against the heart and an alternating current of 1.5 amperes successfully defibrillated the ventricles and saved the patients. Procedures applicable to the surgeon's operating table are vastly different from those encountered in most accidents. The brain and nervous systems remain viable from three to five minutes after circulation ceases. This time interval is so short that once ventricular fibrillation occurs in man, little can be done to save the victim.

The hazard due to direct current is of increasing importance due to the greatly increased use of d-c welding equipment and high-voltage power supplies. The ratio of average d-c release to 60-cycle let-go currents is about 4.8 to 1. Ferris and associates determined the direct current required to produce ventricular fibrillation in 11 sheep for 3-second shocks. The ratio of average d-c to a-c fibrillating current was 5 to 1. This information, although obtained for only 11 animals, is the only quantitative data available on fibrillating direct current. It is to be expected that the fibrillating current for very short shocks would approach the 60-cycle crest value. This is a ratio of $\sqrt{2}$ to 1. If we assume that the lines of Figure 6 apply for durations as short as one fourth cycle, that is, 0.0042 second, a second value may thus be established. A series of straight lines for direct current shocks similar to those of Figure 6 could be drawn, using the factors 5 and 1.41 for shock durations of 3.0 and 0.0042 second, respectively.

The danger from capacitor discharge is of vital importance. It generally is believed that the initial current and charge are of prime importance in determining the hazard. The time constant of the circuit and the stored energy also may be important, but no data are available on this phase of the subject. An idea of threshold fibrillating currents may be obtained from Figure 6, if we assume that the 60-cycle response holds for shocks of 1/4-cycle duration. The one half of one percentile value for man corresponding to 0.0042 second is 2,600 rms milliamperes. Multiplying by $\sqrt{2}$ gives 3.7 crest amperes, which may be considered as the equivalent capacitor initial current.

Table I. Impulse Tests on Animals

No. of Animals	Age, Mos	Est. Weight, Lbs	Open Circuit Voltage	Capacitor Charge, Millicoulombs	Animal Resistance, Ohms	Avg. No. of Shocks	Remarks
Spring lambs. Contacts—metal electrode on lips to rear feet in bucket of salt water							
10	4–6	72	1,400	18.2	550–1,750	3	
10	4–6	70	1,600	20.8	550–1,750	3	Three sheep stunned for 5–15 sec*
13	4–6	72	1,750	21.0	550–1,750	3	One sheep stunned for 5–15 sec*
6	4–6	72	1,750	22.7	550–1,750	2	One sheep stunned for 5–15 sec*
Pigs. Contacts—insulated feed trough to rear feet on wet earth							
1		180–200	1,400	4.2		4	
25		180–200	1,400	7.0		4	
1		215–225	1,000	10.0		1	
2		215–225	1,200	12.0		1	
1		215–225	1,600	16.0		1	
25		215–225	1,750	17.5		1	
1		215–225	1,400	16.8		1	
12		215–225	1,750	21.0		1	One pig stunned f 60 sec*

* Animals were stunned on third or fourth consecutive shock.

Reference 5 contains an account of studies made by the author in the attempt to determine the hazard due to impulse currents. The investigation was made to determine the hazard of capacitor-discharge electric-fence controllers and included a survey of human accidents on lightning generators. A few capacitor discharge tests were made on male volunteer subjects on voltages up to 1,750. Impulse tests using larger capacitors were made on spring lambs and pigs. Table I shows some of the data obtained from tests made on the animals. No fatalities were experienced and higher power was not used because of limitations of equipment. No attempt was made to co-ordinate the shocks with the sensitive phase of the heart-cycle. However, because of the large number of shocks applied, it is reasonable to assume that they were distributed throughout the heart-cycle, and probably many occurred during the sensitive phase. It would seem that the highest currents given the smallest animals might give some idea of reasonably safe upper limits. The average weight of the spring lambs was 72 pounds or 32.6 kilograms, maximum voltage 1,750, and maximum charge 22.7 millicoulombs. Considerable difficulty was experienced in measuring the body-resistance of the animals because of unavoidable variations in contact resistances. Best results were obtained using a d-c ohmmer, the values ranging from 550 to 1,750 ohms. Using a conservative value of 930 ohms for circuit, animal, and contact resistance, equation 6 gives an initial current of 3.7 amperes which compares with that suggested in the preceding paragraph. If we assume that equation 6 holds for both current and charge, the corresponding value for man is 45 millicoulombs.*

The criterion proposed for defining dangerous current thresholds is based on the current which might produce ventricular fibrillation in one half of one per cent of a large group of normal adult men. The choice of course is arbitrary, and any other number could be selected. There is no intention to be ruthless about the remaining one half per cent; however, one has to stop somewhere, and the 1/2 percentile was considered a reasonable stopping point. Perhaps the best justification for choosing the 1/2 percentile is that it was used in estimating reasonably safe let-go currents and provoked no adverse comment.[1,2,3] As previously mentioned, all available experimental data were analyzed carefully, and in instances which required special judgment (such as just where to draw the best line to determine the general trend of a response), an attempt was made to adopt procedures which would give a conservative final result. Several different attempts were made to correlate the data, but the method proposed is believed to be the best that can be done. It is indeed unfortunate that statistical predictions must be expressed numerically, as there is a tendency to place too much emphasis on mere numbers. Quite aside from the uncertainties of the present analysis, there remains the important uncertainty of transferring results obtained from animal experimentation to man. However, data from numerous human accidents, although meager and inconclusive, seem to

* The charge or quantity of electricity passing a given point in a circuit is given by:

$Q = 10^{-3}\int i\,dt$ millicoulomb (milliampere-seconds)

For a capacitor of C microfarads charged to E kilovolts,

$Q = CE$ millicoulomb (milliampere-seconds)

An Electronic Drive for Windup Reels

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REELING OPERATIONS impose a number of special requirements on the reel drive, particularly if the latter is of the so-called core type as compared to the rather rarely used peripheral type drive.

In a core type drive the reel is driven through a center axle on which it is mounted. This type is usually preferred to the peripheral type drive which involves traction along the periphery of the reel and thus requires a direct contact and a sufficiently high tangential force exerted upon the material being wound. It is apparent that, although the peripheral drive usually will represent a much simpler electric system since it is not influenced by the change in the diameter of the reel, it is often undesirable or even impossible to transmit the driving torque directly through the material involved in the winding operation.

The core type drive, on the other hand, does not present these disadvantages, and is used most commonly, although it normally involves other problems and the resulting complications of the electric system.

Two basic quantities usually are associated with any reeling operation where a continuous strip of material such as yarn, fabric, wire, or steel, for example, is drawn by the so-called take-up reel and wound on it, so that the diameter of the reel gradually increases during the process. The first quantity is the speed, and the second, the tension of the strip. Some reel systems impose exacting requirements as to the automatic regulation and control of one or both of these quantities.

This paper presents the description of a recently developed electronic drive and control for a core type take-up reel where the speed of the strip is maintained constant automatically during the entire reeling operation. In addition, the speed of the strip can be preselected by the operator within a wide and stepless range. The problem of tension of the strip is not considered here. In many of the applications, tension of the strip is of secondary importance, and proper control of speed is a fundamental requirement. Furthermore, tension regulating devices always can be added, if so required, by proper control of the unwinding reel. The drive described in this paper has been developed for take-up reels in systems where the reel also performs the drawing of the strip and thus determines the speed of the line.

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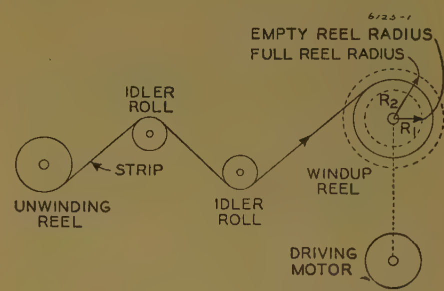

Figure 1. Outline of a windup reel system

Principles of Core Type Reeling

In Figure 1 is shown schematically a take-up reel with its minimum and maximum diameters. The radius of the reel builds up during the reeling operation from a minimum value R_1 to a maxi-

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indicate that the proposed threshold values are conservative. It is the author's opinion that currents much in excess of the proposed threshold values must be considered very dangerous to human life. Perhaps, at some future date, sufficient time and funds may be available and a more accurate solution be obtained.

We are indebted to H. B. Williams for the following discussion of effects at high frequency. On sinusoidal high-frequency alternating currents, or on repeated current pulses of very short duration, account must be taken of the fact known to physiologists that, as the shock duration decreases, its strength must be increased in order to produce the same stimulation. As the duration becomes very small, this increase must be very great, finally becoming so great that destruction of living substances may occur before it can respond. At higher frequencies, large currents may pass without causing stimulation of muscles or nerves, and these may cause deep heating effects. Since the heat-sensitive mechanism is located in the skin, there is a possibility of damage to internal organs by high-frequency currents even though no very unpleasant sensations may be apparent. The currents necessary to produce this effect would be in the order of an ampere or more. High-frequency currents of several hundred milliamperes are used quite commonly by the medical profession for deep heating. This form of treatment is called medical diathermy.

In concluding, it should be mentioned that, because of the wide variation in the physical condition of individuals, an occasional death is to be expected from casual contact involving electric currents known as safe for most normal, healthy individuals, but these are probably not fibrillation deaths. It is gratifying indeed that victims surviving the initial shock of an electrical accident seldom suffer serious aftereffects or other permanent disability. Since it is impossible for the layman to distinguish between respiratory inhibition, ventricular fibrillation, and heart failure, he should begin artificial respiration immediately and summon medical aid as soon as possible.

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mum value R_2. The speed of the strip may be expressed by the following equation

$$v = 2\pi nR \text{ feet per minute} \tag{1}$$

where

n = motor speed in revolutions per minute
R = instantaneous radius of the reel in feet

The torque developed by the motor

$$T = FR + T_0 \text{ pound-feet} \tag{2}$$

where

F = tension of the strip in pounds
T_0 = torque loss due to friction in the reel, gears and windage

The power developed by the motor may be expressed as

$$P = 2\pi nT$$

or, considering equation 2,

$$P = 2\pi n(FR + T_0)$$
$$P = 2\pi nRF + 2\pi nT_0 \tag{3}$$

The first term of equation 3 obviously represents the useful power required to draw the strip, and is the product of the strip speed v, and strip tension F. The second term, which is the product of the motor angular speed and the friction torque of the reeling system at the motor shaft, represents the loss of power of the reeling system. Thus, the efficiency of the reeling system

$$\eta = \frac{1}{1 + \dfrac{T_0}{RF}} \tag{4}$$

It is apparent from equation 1 that in order to maintain constant line speed v, the speed of the driving motor must change in inverse proportion to the diameter of the reel. On the other hand, if the friction torque T_0 is neglected, then the power required during the reeling operation

$$P = vF = 2\pi nRF = \text{constant}$$

Thus, if a constant line speed is to be maintained during the reeling from an empty reel with a radius R_1, to a full reel with radius R_2, the speed of the driving motor n corresponding to any intermediate radius R of the reel should be

$$n = n_1 \frac{R_1}{R} \tag{5}$$

where n_1 is the motor speed for an empty reel.

As was mentioned previously, the power requirement imposed on the drive during reeling remains essentially constant and, consequently, the most adequate and economical way of obtaining the required speed control is by controlling the field excitation of a d-c shunt-wound motor.

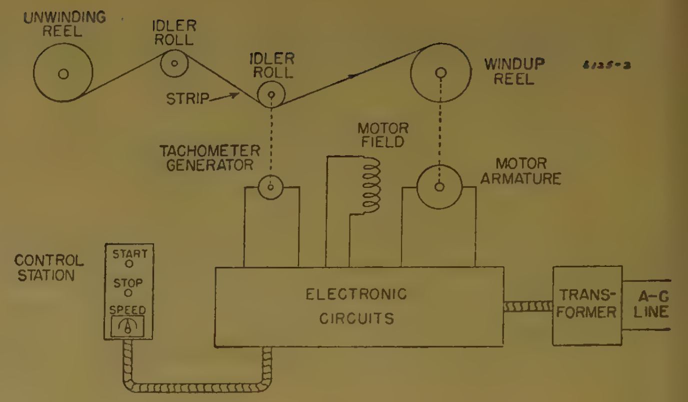


Figure 2. General scheme of the electronic drive for windup reels

Another type of speed control is necessary to provide an adjustable strip speed, and it is often required that a speed range as wide as ten to one be provided by the drive. Here, if an approximately constant tension is assumed, for any given radius of the reel, the required torque is independent of the line speed (see equation 2), and the required horsepower is directly proportional to the line speed v (if the friction torque T_0 is neglected). Consequently, the armature voltage control of a d-c shunt-wound reel motor is the most suitable means of obtaining the control of the strip speed.

A clear distinction between the two types of the reel-motor speed control should be specifically emphasized. For any given diameter of the reel, the line speed should be adjustable by armature voltage control of the driving motor, and the speed of the motor should remain constant for a given excitation and a given speed setting, even if the torque is varied within wide limits. This speed control should not respond to changes in the diameter of the reel.

The second type of speed control should respond automatically to any change in the diameter of the reel, so that with increasing diameter, the excitation of the motor should increase automatically to slow the motor down just enough to prevent the speed of the strip from being increased. The field control of the motor should not respond, generally speaking, to any change in line speed obtained through the armature voltage control.

Basic Control System

To maintain a constant strip speed with varying diameter of the reel, an electronic field regulator is incorporated into the system. The term regulator, or servomechanism means an automatic control system characterized by a closed cycle of interdependence of quantities involved, so that a slightest deviation of the regulated quantity from the predetermined value results in the appearance of the so-called restoring force, which acts in the direction to oppose any change in regulated quantity, or, if some change has already occurred, to restore the quantity to its original value. Actually, the most accurate and sensitive regulator is unable to restore the deviated quantity to exactly its original value, when a definite cause for the deviation exists. This is so because the regulator creates a restoring force or "back-to-normal" action only in response to some existing deviation, and a new balance of the system always will be established only with the existence of some deviation. However, a satisfactory regulating system will reduce the deviation to such a minimum that, depending upon the application, it may be assumed to be negligible.

Any regulator is characterized by the following four basic elements which form the links of the previously mentioned closed cycle of interdependence of quantities:

1. Indicator of the regulated quantity.
2. Standard reference quantity with which the regulated quantity is compared.
3. Amplifier which originates a sufficiently strong restoring force even for minor deviations of the regulated quantity from the prescribed value.
4. Restoring force which manifests itself in an action opposing any deviation of the regulated quantity.

In addition, a regulator may have a number of secondary elements such as damping (antihunting) and anticipatory elements which play a vital part in the stability and the time of recovery of the regulating system.

The electronic reel drive is shown schematically in Figure 2. A tachometer generator driven by the strip through an idler roll is used as the indicator of the regulated quantity, that is, of the line speed. Since this generator is required to deliver only negligible

power, its size may be small and, consequently, its WR^2 and losses equally small. The preferable tachometer generator is a d-c permanent-magnet type, as here there is no need to provide separate excitation or, as in the case of an a-c type, to amplify and then rectify the output to make it a useful indication of speed. Besides, the permanent-magnet type tachometer generators usually have an excellent voltage-speed characteristic. The output voltage of a d-c permanent-magnet tachometer generator represents a linear function of its speed, particularly when the current output is negligible. The voltage output is relatively high, of the order of 50 volts per 1,000 rpm, and for all practical intents and purposes, independent of temperature changes or any other external influences. A typical voltage-speed characteristic of a permanent-magnet tachometer generator is shown in Figure 3.

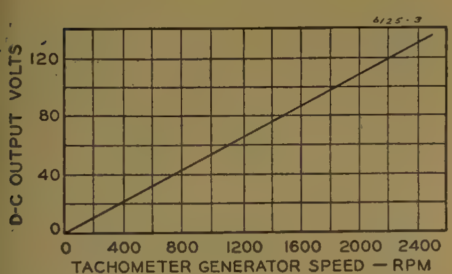


Figure 3. Typical output voltage of a d-c permanent magnet tachometer generator

The output voltage of the tachometer generator (see Figure 2) is introduced into the field control portion of the electronic drive where it is balanced or compared with a reference voltage. The difference of the two is then fed into an amplifier, and after the amplification, is used to control the output of a thyratron-exciter which supplies the field current to the reel motor. As it will be seen later, the control circuits are so designed that when the line speed tends to increase because of the increase in the diameter of the reel (see equation 1), the field of the motor is strengthened to slow the motor down in accordance with equation 5. This forms the previously mentioned closed cycle of interdependence of quantities, which is represented in Figure 4.

The armature power supply and armature control of the electronic drive is obtained by means of a controlled rectifier consisting of thyratron elements and various control circuits. These circuits provide a constant-speed characteristic so that for a given excitation of the field, the speed of the motor remains essentially constant for any variation of torque within the operating range of torque values. The armature control circuits are of the voltage regulated, *IR* drop compensated type. They provide a properly rising voltage-current characteristic and essentially constant speed-current characteristic. Current limit features also are incorporated in the armature control circuits, to take care of the starting and overload conditions of the drive.

In addition to a constant speed-torque characteristic, the armature control circuits provide means for adjustment of the motor speed by voltage control, and in that manner the strip speed can be changed. Each setting of the speed control dial corresponds to a different but strictly determined line speed, automatically regulated to stay constant and independent of the changes in the diameter of the reel. Usually, the line speed control dial is a small $1^1/_2$ inch in diameter potentiometer, assembled as a part of a separately mounted control station, which also includes other operator's controls such as start and stop push buttons and threading speed push button. The operator's control station usually is mounted remotely from the cabinet housing all the electronic controls and at a location most convenient for the operator. The armature control portion of the drive as well as the operator's control station is shown schematically in Figure 2.

Special Problems

Outside of system stability problems which are, of course, present in any regulating system and require special stabilizing means, the following two problems are of particular importance:

1. Field control during the acceleration.
2. Field regulator response to a change in line speed resulting from the readjustment of that speed by the operator.

Although these problems will be discussed in more detail later, in connection with the description of circuits, it seems appropriate to mention their nature in advance.

Referring to the acceleration problem, it should be noted that at the instant of starting the reel drive, the voltage output of the tachometer generator is zero. Furthermore, at the start of the reeling operation the reel is empty and, obviously, the driving motor and reel should accelerate to a maximum operating speed, corresponding to the lowest operating value of the motor field. Consequently, it follows that on starting of the drive, when the speed of the line is zero, the regulator, responding to a low speed signal of the tachometer, will naturally tend to decrease the field excitation of the motor to a minimum. Considering a relatively low value of starting current limited by the current limit circuit (see section on armature control) to about 200 per cent of rated armature current, it will become apparent that a very low starting torque will be developed by the reel motor so that the latter may not accelerate at all. To prevent this, a special electronic accelerating circuit is added to the field regulator. When the reel motor is not running, this field accelerating circuit maintains full field current, and when the motor is started, the field still is maintained during the initial portion of the accelerating period. Then the field current gradually is decreased exponentially with time, but, owing to a special electrical coupling between the armature and field control circuits, it never is allowed to decrease below a certain minimum value, as long as the armature current exceeds the rated value more than by about 50 per cent. As soon as the armature current drops to its normal operating value, indicating that the acceleration of the motor is completed, the action of the field accelerating circuit disappears, and the field current of the motor is controlled by the regulator through the tachometer generator which has by that time accelerated to its operating speed. From now on the regulation of the line speed proceeds as described previously.

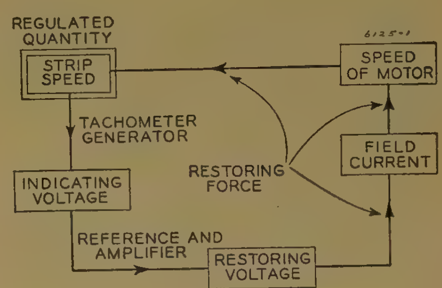


Figure 4. Cycle of interdependence of the strip speed regulator

The problem of the field regulator response to a change in tachometer generator signal, when caused by the readjustment of the line speed by the operator, deserves special attention. As pointed out before, the slightest change in the tachometer generator output voltage, indicating a change in line speed, normally

will result in the action of the regulator to properly change the motor excitation and its speed in such a way as to oppose any change in the tachometer generator signal voltage. However, if the change of the tachometer generator voltage is caused by the line speed change through armature voltage control, the regulator obviously should not respond to that change. In other words, when the change in the tachometer generator voltage is a result of the change in diameter of the reel, the regulator should respond and control the motor field, but for any given diameter of the reel, the field current of the motor should remain essentially the same for different line speeds.

This requirement is fulfilled by using two potentiometer units mounted in tandem as operator's line speed control. One of the potentiometer units is used in a regular manner as the armature speed control, the other properly changes the field regulator reference voltage. Thus, for example, when the knob of the speed control is operated to call for a higher line speed, the tachometer generator signal voltage will become higher, and this normally would cause the regulator to increase the excitation of the motor. However, the operation of the control knob in the direction to increase the line speed also will increase the reference voltage of the field regulator in such a manner that the difference between the two voltages will remain the same, and the field current will not be changed.

The voltage across the reference potentiometer of the regulator can be easily adjusted so that a change in line speed and in armature voltage is accompanied by a proper change in reference voltage of the regulator. In fact, this adjustment by no means is critical, and some small changes in the field current for different operating line speeds can be tolerated, since this will not produce any adverse effect on the operating or regulating properties of the drive.

Electronic Circuits of the Drive

A simplified schematic diagram of the electronic drive is shown in Figure 5. It includes all the elements necessary for the proper explanation of the principles of operation of the system. For the sake of simplification, details of various auxiliary d-c power supplies have been replaced by batteries. Heater circuits for all the electronic tubes have been omitted, and all the grid controlled tubes, both of the thyratron and high vacuum type, are represented as triodes. All the by-pass and filtering capacitors also have been omitted. The diagram covers the entire electric system of the reel drive and is divided distinctly into the armature control and the field control portions.

Armature Control Circuits

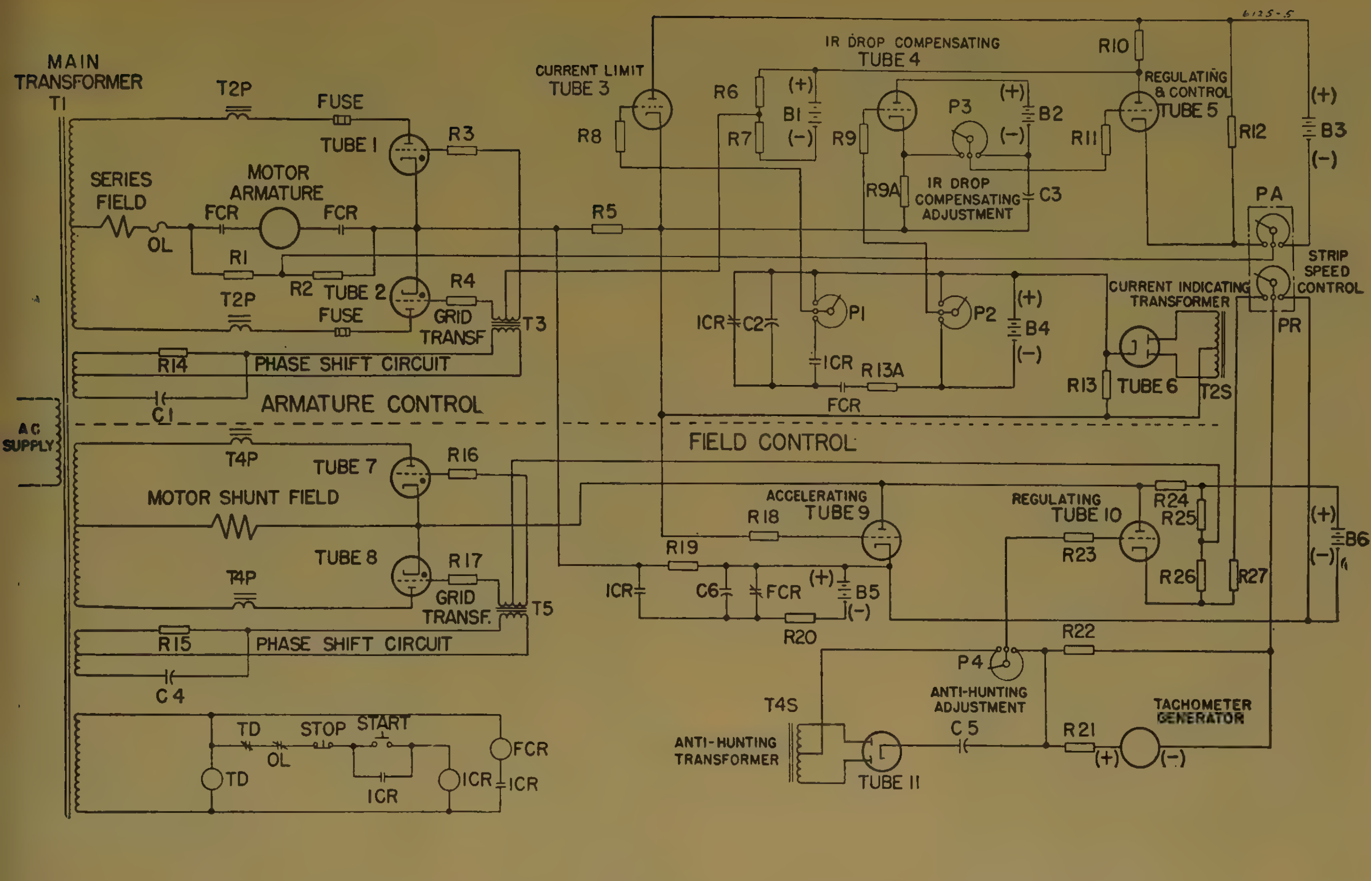
The armature control circuits are similar to those described and analyzed in detail in a previous paper.[1] They consist, as shown in Figure 5, of four major groups: the controlled power rectifier, master regulating and control circuits, *IR* drop compensating circuits and the current-limit circuits. The power rectifier circuit is shown as a symmetrical 2-phase rectifier (often called single-phase full-wave rectifier) although it may be also a 3-, 4-, or 6-phase unit, depending upon the horsepower rating of the drive. The rectifier consists of a center-tapped winding of the main transformer *T*1, two thyratron tubes 1 and 2, contacts *FCR* of the main contactor, fuses in the anode circuits, grid transformer *T*3, and current transformer *T*2. A phase-shift circuit consisting of a center-tapped winding of transformer *T*1, resistor *R*14, and capacitor *C*1 provides a voltage which lags the rectifier anode supply voltage by about 90 to 100 degrees. This lagging voltage is applied to the primary winding of the grid transformer *T*3. The grid control voltage of the thyratrons consists of an a-c component supplied by transformer *T*3 and a d-c component. The latter is applied between the cathodes of tubes 1 and 2, and the center tap of the grid transformer *T*3. During the control process the angle of ignition of the thyratrons is changed either automatically or manually by proper changes in the d-c component which may assume both negative and positive values so that the constant a-c component of the grid voltage, superimposed upon the variable d-c component, can be moved up or down and control the angle of ignition in a well-known manner.

Control of the d-c component of the thyratron grid voltage is obtained through the variable voltage across the load resistor *R*10 of the main control tube 5. If tubes 3, 4, and 6 are disregarded temporarily, it will be easy to notice that the armature system is an adjustable voltage regulator. Thus, the voltage across *R*2 is the indicator of the regulated quantity, being proportional to the armature voltage. The adjustable portion of the potentiometer *PA* represents the reference voltage. The *PA* voltage is negative with respect to the grid of tube 5 and the *R*2 voltage is positive. Since the voltage across the adjustable portion of *PA* always is higher by several volts than the voltage across *R*2, their difference applied to the grid of the master control tube represents a negative voltage of several volts.

The grid circuit of the thyratrons is completed from the cathodes through *R*2, *PA*, *R*12, *R*10, and *R*6 to the center tap of transformer *T*3. All the d-c components in that portion of the circuit essentially are constant except the voltage across *R*10 which changes with the plate current of tube 5, and performs the actual control of the angle of ignition of the power thyratrons. It is important to note that the direct voltage across *R*10 has a negative polarity with respect to the grids of thyratron tubes. Thus, when the plate current of tube 5 is high, the angle of ignition of the thyratrons is delayed greatly, and if the current of tube 5 still will increase, the thyratrons will be cut off altogether. Conversely, when the plate current of tube 5 is decreasing, the negative voltage component across *R*10 will decrease, and the angle of ignition of the power thyratrons will be advanced, increasing either the voltage or the current output of the armature rectifier, or both.

It is apparent that the voltage across *R*10 together with the grid control of the thyratrons represents the restoring force of the armature voltage regulating system, since any deviation of this voltage from the prescribed value will result in a response of the amplifier 5 and a change in voltage across *R*10 in the direction to oppose the deviation of the armature voltage. Furthermore, a change in the reference voltage obtained by the readjustment of *PA* will result in a change in rectifier voltage output so that a new balance of the system will be obtained at a different armature voltage, that is, a different speed of the motor. The difference between the reference and the indicating voltages, however, will not be changed appreciably because of the strong amplifying action of the regulating tube 5.

It is well known that if the armature voltage remains constant, for a constant excitation, the speed of the motor usually will drop with increasing torque, as is normally the case for shunt-wound motors. This drop is caused mainly by the *IR* drop in the armature circuit. Although this drooping speed–torque characteristic is usually not objectionable at rated motor speed, at lower speeds such as one-tenth of the base speed, a



normally drooping characteristic of about 10 per cent will mean a very poor speed regulation and even the inability of the drive to develop full torque.

In order to correct this objectionable droop, an *IR* drop compensating circuit is provided. The main element of the compensating circuit is tube 4, which is controlled automatically by a grid voltage proportional to the armature current. The indication of the latter is obtained through a special current transformer *T*2 whose two primary current windings *T2P* are connected in series with the anodes of the power thyratrons 1 and 2. Although each of the windings conducts unidirectional current pulses, connections are such that the resultant magnetic effect is the same as if an alternating current were flowing in the current windings. This prevents the saturation of the current transformer. The voltage generated in the voltage winding *T2S*, generally speaking, is proportional to the motor current. This voltage is rectified by means of tube 6 and then filtered out (the filter is not shown in Figure 5) so that the direct voltage across *R*13 may be used as an indication of the armature current.

There are two main voltage components in the grid circuit of tube 4: the negative bias voltage obtained from *P*2, necessary to determine the proper operating point of tube 4, and the positive current-

Figure 5. Schematic diagram of the electronic drive for windup reels

indicating voltage, described in the preceding paragraph. Thus, an increase in motor torque and motor armature current will cause an increase in plate current of tube 4 and an increase in magnitude of the voltage across *P*3. It will be noted that *P*3, which represents the *IR* drop compensating adjustment, forms part of the grid circuit of the master control tube 5. The voltage across the adjustable portion of *P*3 forms an additional negative component of the grid voltage of tube 5, and with increasing load current of the motor, the grid of tube 5 becomes more negative, the plate current decreases, and the firing point of the power thyratrons is advanced just enough to compensate for increasing *IR* drop of the motor. Conversely, the opposite will occur with decreasing load of the motor.

Tube 3 acts as a current-limiting element of the armature control system. This tube, which is characterized by a sharp cut-off, is biased off by an adjustable negative voltage obtained from *P*1 and does not conduct any current for all the normal loads of the motor. However, when the load current of the motor exceeds a certain critical value, the voltage across *R*13 becomes so high that tube 3 will start to conduct. It will be noted that the load resistor of tube 3 designated as *R*5 is in the grid circuit of the master control tube 5. As soon as tube 3 starts to conduct, a voltage appears across the resistor *R*5, and the polarity of this voltage is positive with respect to the grid of tube 5. Since tube 3 itself provides a considerable amplification, a very small increase of the voltage across *R*13 beyond its critical value results in appearance of a much higher voltage across *R*5. As a result, the plate current of tube 5 is increased, the firing point of the thyratrons delayed, and the armature voltage lowered.

If the motor current continues to rise, the delay of firing point of the rectifiers will soon become so great and the armature voltage so low that the motor will stall with the armature current reaching its limit value. This limit is normally of the order of 200 per cent of the motor rated current, although it may be adjusted by means of the potentiometer *P*1. The current limit which, for a given excitation of the motor, may be called torque limit, plays an obviously important part during the acceleration and overload of the drive.

Capacitor *C*2 connected in parallel with the current limit control potentiometer *P*1, as well as contacts 1*CR* and *FCR* in series with *P*1, have an important function during the starting and stopping

of the drive. On stopping of the drive, contact $1CR$ in series with $P1$ opens first and causes an immediate disappearance of the negative bias voltage for tube 3, so that the latter will conduct its full current. This, of course, will cause the tube 5 also to conduct full current, and the thyratron tubes will be cut off, even before the FCR contacts will have time to open completely. As a result, no arcing across main FCR contacts will occur.

When the motor is stopped, capacitor $C2$ is discharged by a parallel contact of $1CR$. On starting the motor, the thyratrons are cut off by grid control as initially there is no voltage across $P1$. After the closure of the relay $1CR$ and contactor FCR, the voltage across $P1$ appears gradually since it has to follow the voltage across capacitor $C2$ which is charged from the source $B4$ through the resistor $R13A$. Consequently, the firing angle of the thyratrons is advanced smoothly until the current limit circuit takes over the control of the acceleration. Thus, during the starting of the drive, no objectionable current surges may occur in thyratrons 1 and 2, and no sparking on the main contacts of FCR ever takes place.

Field Control Circuits

A general outline of the strip speed regulator operating through the field control of the motor was given in one of the previous sections of the paper. A simplified schematic diagram of the strip speed regulating system is shown in Figure 5. It must be noted that the power rectifier portion of the electronic field exciter is fundamentally the same as the power rectifier for armature control, described in the previous section. It consists of two thyratron rectifiers 7 and 8, a center-tapped transformer winding, a grid transformer $T5$, and a phase-shift circuit which provides the constant a-c component of the thyratron grid voltage. The current transformer $T4$, however, has a function different from that of transformer $T2$.

The variable d-c component of the thyratron grid voltage is developed between the cathodes of the thyratrons and the center tap of the grid transformer. It consists of two voltages: a constant negative voltage across $R25$ and a variable positive voltage across $R24$. The voltage across resistor $R24$ is changing with the plate current of either tube 10 or 9, or both.

Tube 10 is the field regulating tube and represents the amplifier of the strip speed regulator. The grid circuit of tube 10 includes the negative reference voltage across $R27$ and a portion of PR, and the positive strip-speed indicating voltage across $R22$ obtained from the tachometer generator. The reference voltage always is higher by several volts than the indicating voltage, and the difference of the two represents a negative voltage of several volts applied to the grid of tube 10.

It should be noted that the reference voltage in Figure 5 is obtained as part of the main voltage divider which also includes resistors $R25$ and $R26$ and is supplied from the d-c source $B6$. Although the plate current of tube 10 will flow through $R27$ and PR and will change the reference voltage, such a simplified arrangement is permissible if the resistance of $R27$ and PR is relatively low so that the voltage drop in these resistive

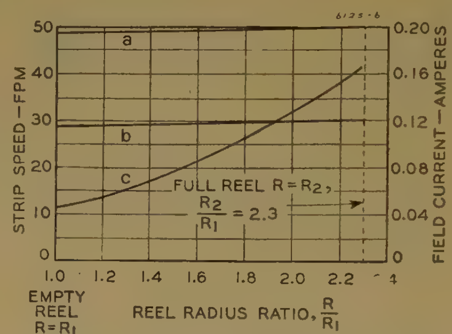


Figure 6. Typical operating characteristics of the electronic drive

elements caused by the plate current of tube 10 is small. Furthermore, one must remember that during the operation of the regulator the plate current of tube 10 changes only slightly because of the general amplification of the system, so that the actual change in reference voltage caused by the change in plate current of tube 10 is negligible.

As an example, with total reference voltage of 25 volts, the resistance of $R27$ and PR equal to 1,000 ohms and the change in the operating plate current of tube 10 of 0.1 milliampere for the entire reel-diameter build-up of two to one, the resulting change in reference voltage will be equal to 0.1 of one volt. Thus, the error caused by the change in the reference voltage will amount to 0.4 of one per cent.

Although the figures cited in the previous paragraph are based on actual test records, they may vary for different motors and different reel-diameter ratios, and in some cases it may be advisable to provide a separate d-c supply for the reference voltage of the regulator, particularly if extreme accuracy of the regulator is required.

The general principles of operation of the regulator were described in one of the previous sections. With reference to Figure 5, it is apparent that when the speed and the signal voltage of the tachometer generator tends to increase because of the increasing radius of the reel, a slight increase in voltage across $R22$ will cause the grid of tube 10 to become less negative, and the plate current of this tube will increase. This will cause the appearance of a restoring force which will manifest itself through an increase in voltage across $R24$, the advancement of the firing point of the thyratron tubes 7 and 8, and the resulting increase in the excitation current of the reel motor.

If the strip speed is increased by a clockwise turn of sliders of the combination unit PA–PR, the tachometer generator voltage as well as the voltage across $R22$ will increase in proportion with the strip speed, but if properly adjusted, the reference voltage across the active portion of PR will increase likewise. In that manner, the difference of the two voltages in the grid circuit of tube 10 will remain the same and the field current of the reel motor will not change.

Under normal operating conditions, during the regulating process, the accelerating tube 9 does not conduct any current being biased off by the negative grid voltage across $R19$. It will be noticed that in addition to resistor $R19$ the grid circuit of the accelerating tube 9 also includes the resistor $R5$ whose significance in the armature control circuits was described previously. This resistor, however, plays a very important part in the field control circuits as well.

Before the drive is started the relay $1CR$ is de-energized, and there is no voltage across the resistor $R19$. Consequently, the accelerating tube 9 conducts full current resulting in full voltage across $R24$ and full exciting current of the reel motor. After the start pushbutton is depressed both $1CR$ and FCR pick up and the negative bias of tube 9 gradually appears across the resistor $R19$. The increase of the voltage across $R19$ is gradual because of the charging of capacitor $C6$ through resistor $R20$. As a result, the plate current of tube 9 and the voltage across $R24$ will decrease slowly and so will the excitation current of the motor. In the meantime, however, the motor and the tachometer generator will continue to accelerate, so that at a

certain moment when the tachometer generator voltage becomes high enough, the regulating tube will start to conduct current and will take over the control of the motor excitation in the manner as described previously. If, however, the motor does not have enough time to accelerate to its operating speed before the capacitor $C6$ is charged completely, the presence of the positive voltage across $R5$ will prevent both the plate current of tube 9 and the field current of the motor to decrease below a certain minimum value, so that an adequate accelerating torque will be maintained. As soon as the acceleration of the reel system is completed, the armature current will drop to a normal operating value, the voltage across $R5$ will disappear, and, if the charging of capacitor $C6$ was completed before, the plate current of tube 9 will disappear also.

The antihunting circuit of the regulator consists of transformer $T4$, tube 11, capacitor $C5$ and potentiometer $P4$. Transformer $T4$ is basically of the same type as transformer $T2$ which was described previously. The voltage obtained in winding $T4S$ of this transformer is rectified by means of tube 11, but the d-c component of the rectified voltage does not appear across $P4$, being blocked by capacitor $C5$. However, if the field current will tend to change abruptly, a transient current caused by charging or discharging of capacitor $C5$ will flow through $P4$. Thus, a transient voltage will appear across $P4$ and this voltage always will be of such polarity with respect to the grid of tube 10 that an action opposing a fast change in the restoring force will occur. It should be noted that this damping action will take place in response to the rate of change of the average field current and not to the change itself. Thus, if the average field current tends to increase abruptly, a high charging current of capacitor $C5$ will cause the appearance of a transient negative voltage in the grid circuit of tube 10. This voltage will tend to decrease the plate current of the regulating tube, and, consequently, oppose the previously mentioned rise of the field current.

Figure 7. Main elements of a 3-horsepower drive for windup reels

Performance and Conclusions

Typical operating characteristics of the electronic drive are shown in Figure 6. They represent the strip speed in feet per minute as a function of the reel-radius ratio for two different strip speed settings.

The corresponding change in the field current of a 3/4-horsepower reel motor is also shown in Figure 6.

It will be noted that in the particular case represented in Figure 6 the radius of the reel increased 2.3 times during the entire reeling operation. Theoretically, if a constant speed motor were used (an induction motor for example) the strip speed during the reeling would increase in proportion, that is, the speed at the end of the reeling operation would constitute 230 per cent of the speed at the start of reeling.

With the electronic drive the strip speed changed from 48.8 feet per minute for empty reel to 50.0 feet per minute for a full reel, that is, increased by 2.46 per cent. The accuracy of the regulator is related closely to the general stability of the whole system and depends upon the type of the motor, the reel-diameter ratio, the behavior of the back-tension during the reeling operation and other similar factors. For many systems of the more stable character, the accuracy of regulation can be easily improved to one or even one-half of one per cent at full strip speed. For lower strip speeds the amplification of the system will be lower, and the percentage of regulation will be increased correspondingly. Thus, for a strip speed set at 30 feet per minute (see Figure 6), the speed will change from 28.8 feet per minute to 30.0 feet per minute or by 4.16 per cent.

An important condition for a satisfactory operation of the drive is the selection of a proper operating range of the motor field current. The portion of the motor magnetic characteristic which represents the range of appreciable saturation should not be used, since over that range the overall amplification of the system is reduced considerably. It may be noticed from the field current graph in Figure 6 that the range of high saturation of the motor has been left outside the operating range of the system. In other words, if the motor is designed so that its full field current corresponds to an appreciable saturation of its magnetic characteristic, the operating full field current should be sufficiently below the rated full field current of the motor.

A typical 3-horsepower electronic drive for a windup reel is shown in Figure 7. It includes the complete electronic rectifier and control system enclosed in the cabinet, the reel motor, the tachometer generator and the operator's control stations. In addition to the parts shown in Figure 7, a complete system also includes a power transformer, not shown in the figure.

Reference

1. Electronic Control of D-C Motors, K. P. Puchlowski. AIEE Transactions, volume 62, 1943, pages 870–7.

Protection of Industrial Plants Against Insulation Breakdown and Consequential Damages

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Synopsis: Experience has shown that multiple electrical failures may occur simultaneously in an industrial plant as the result of an initial breakdown of insulation causing high transient voltages. Available methods of surge protection to guard against insulation breakdown and neutral grounding to prevent high transient voltages are presented. Consideration is given to the merits and limitations of these methods when applied to industrial plants.

PROTECTION against insulation breakdown and its consequential damages long since has been considered paramount by utility engineers. Their experiences in supplying dependable power at low cost have proved that in the electrical industry there is much wisdom in the old adage, "An ounce of prevention is worth a pound of cure." As industrial plants become more and more electrified, and they are doing so rapidly, industrial engineers likewise must pay more heed to this very important problem. It is the purpose of this paper, therefore, to review some of the factors involved and to outline modern methods of protection that have proved beneficial in industrial plants.

Fundamental Considerations

Electricity under control is mankind's most submissive form of power and functions with almost magic imperceptibility. However, it is ever on the alert to escape its fetters and, once on the loose, it becomes a violently destructive force that defies submission. It is imperative, therefore, that great care be taken to keep it under control, that its path of destruction be limited in case it gets out of control, and that it be brought back under control as quickly as possible. In the language of the protection engineer, these fundamental requirements are:

1. Guard against insulation breakdown or flashover by adequate surge protection.

2. Prevent high transient voltages or severe burning as a result of insulation failure by proper neutral grounding.

3. Clear faults rapidly by fast and well-co-ordinated relaying.

It is intended that this paper deal mainly with the first two requirements.

Protection Against Insulation Breakdown

The insulation of a circuit or piece of apparatus will withstand safely the normal operating voltage for which it is intended unless it has become defective as a result of moisture, overheating, or mechanical damage. Trouble from these causes can be minimized by proper maintenance, careful handling, and periodic maintenance. However, it is not practical to build enough insulation into distribution circuits or apparatus windings to withstand surge voltages to which they may be subjected. Protection against these surges, therefore, must be provided commensurate with the severity of surges to be expected and the type of insulation involved.

Industrial plants that receive their power from distant sources usually are supplied over one or more high voltage circuits, 33 kv or higher, into transformers that step down the voltage to 13.8 kv or less for distribution in the plant. If the source is close by or within the plant, power generally is supplied directly at distribution voltage. The circuits may be underground, overhead, or a combination of the two. If they are entirely underground they are free from lightning and require no surge protection. If overhead and exposed, they are subject to lightning surges that may flash them over, break down the insulation of apparatus connected to them, or both. Where combination overhead and underground is used, it is possible for dangerous surge voltages to build up at the junction points between the overhead and underground circuits. Surge voltages also may pass on into the cable and reflect to high values at the far end. The type of circuits and methods of apparatus connection, therefore, have a direct bearing upon the overall scheme of protection which should include protection against both line flashover and breakdown of apparatus insulation.

Line Flashover

Overhead lines can be insulated and shielded to be practically free from lightning flashover.[1] While this commonly is done with high voltage circuits, 69 kv and above, where the insulation and clearances are inherently high, it usually is not considered practical for the lower voltages because of the extra insulation and conductor spacings required. There are exceptions, however, where shielding of low voltage lines can be justified to provide protection to important apparatus connected to the line. Such circuits often are shielded partially by near-by buildings, trees, and so forth.

Low voltage lines, 13.8 kv and below, may flashover from induced surges as well as direct strokes. If wood construction is used as it generally is, some of these flashovers do not result in trip-outs, although many do.

It is quite common practice to minimize the disturbance caused by trip-outs by

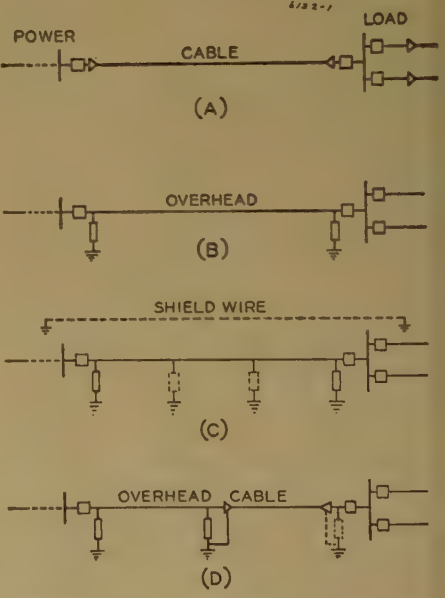

Figure 1. Surge protection of distribution circuits

A—All cable, no protection required
B—Overhead, terminal protection only
C—Overhead, lightningproof protection by either shield wire or spaced lightning arresters
D—Combination overhead, underground. Arresters not required at load end if cable length is less than 100 feet

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rapid reclosing of the feeder breaker. In some cases, where trip-outs are extremely undesirable, line flashovers can be prevented by spacing line-type lightning arresters along the line at 500- to 1,000-foot intervals.

Lightning arresters should be provided at junctions between overhead and underground circuits to protect against high voltages at those points caused by reflected traveling waves. The arresters should be connected directly between the conductors and the cable sheath, and grounded. If the cable is over 100 feet in length, arresters also should be provided at the far end, particularly if the cable feeds into an exposed line or has important apparatus with low surge strength, such as rotating machines, connected to it.

The recommended methods of protecting the different types of distribution circuits that might be involved in an industrial plant are shown schematically in Figure 1.

Apparatus Protection

There are two general classes of electric apparatus that must be considered in the over-all scheme of protecting an industrial plant. One of these includes that equipment, such as oil-insulated transformers, that has surge strengths equivalent to basic insulation levels that have been established for the different voltage classifications.[2] This class of equipment can be protected adequately by standard lightning arresters of the proper voltage rating connected between the exposed line leads and ground. The arresters should be located as close as possible to the apparatus frame or tank as shown in Figure 2.

The second class includes apparatus such as air-cooled transformers and rotating machines that have impulse levels

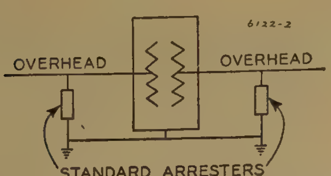

Figure 2. Surge protection of apparatus meeting basic insulation levels

lower than the basic insulation levels. A comparison of the impulse levels of oil-insulated and air-cooled transformers from 1.2 kv to 15 kv classification is given in Table I. The surge strength of rotating machines are normally somewhat less than the values shown for air-cooled transformers. Because of its lower impulse level,

Table I. Comparison of Impulse Insulation Levels

Insulation Class (Kv)	Impulse Levels $1^1/_2$x40 μsec Wave Kv Crest: Oil-Insulated Transformer	Dry-Type Transformer (Proposed)
1.2	30	10
2.5	45	20
5.0	60	25
8.66	75	35
15.0	95	50

this second class of apparatus requires special protection, particularly when connected directly to overhead lines that are exposed to lightning surges.

Air-cooled transformers that are subject to surge voltages should be protected by special arresters, similar to those used for rotating machine protection, at the

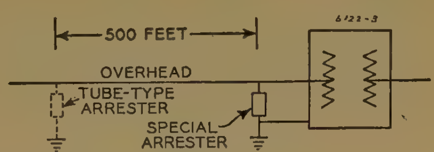

Figure 3. Surge protection of dry-type transformers

transformer terminals,[6] (Figure 3). If the transformer is connected directly to an overhead line, a set of standard arresters, either valve or tube type, also should be connected about 500 feet ahead of the transformer. Tube-type arresters are influenced somewhat less by arrester ground resistance. Liquid-filled transformers should be used in preference to air-cooled transformers where the circuit is exposed highly to severe surges.

Recommended methods of protecting rotating machines are given in a paper by by McCann, Beck, and Finzi.[5] The essential requirements are to limit the magnitude of surge voltage at the machine terminals to protect the insulation to ground, and to slope the front of the surge wave so that it will not pile up across a few turns of the machine winding and overstress the turn-to-turn insulation. The first of these requirements is accomplished by special lightning arresters having low sparkover characteristics, and the second by shunt capacitors and series inductance. These fundamental components of the protective scheme are illustrated in Figure 4*A*. Actually, the series inductance may be a section of the incoming line, 500 feet in length, immediately ahead of the machine (see Figure 4*B*). A standard line-type arrester is required ahead of the series inductance to limit the magnitude of surge voltage that can pass in to the inductance and shunt capacitor combination. This arrester preferably should be a tube type if the ground resistance is more than two or three ohms, because the discharge voltage across the tube collapses to near zero for a half cycle, thus resulting in only a small amount of surge energy being transmitted into the protective devices at the machine terminals. Where the line is exposed to severe surges, or where the arrester ground resistance is higher than about five ohms, one or more additional sets of arresters should be placed out on the line at about 500-foot spacings. The first 500-foot line section ahead of the machine should be shielded (see Figure 4*C*) if there is danger of a direct stroke of lightning in that section.

The special arresters at the machine terminals should be valve type because of its lower sparkover voltage. The best protection is provided by station-type arresters which are recommended where the kilovolt-ampere rating of the machine or machines protected exceeds 1,000. The special line type generally is considered adequate for smaller sized machines.

Where several rotating machines are connected to a common bus, adequate protection can be provided by means of a single set of special arresters and capacitors connected to the bus, if each machine

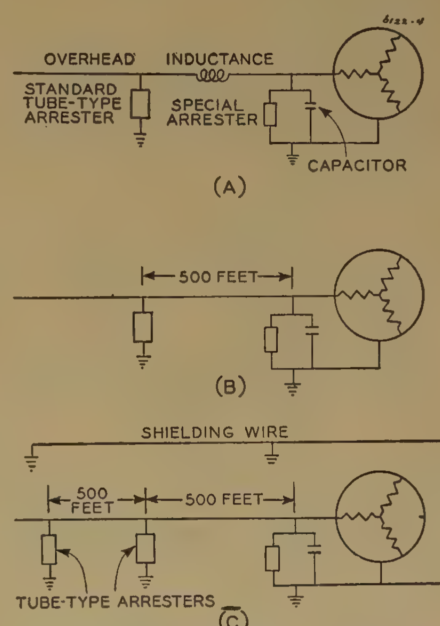

Figure 4. Surge protection of rotating machines

A.—Fundamental scheme
B.—Equivalent scheme
C.—Protection against severe surges including line shielding and spaced lightning arresters

is connected to the bus by an unexposed circuit not over 200 feet in length. Otherwise, each machine should be provided with a set of capacitors and special arresters mounted near its terminals. If there is enough cable connected to the bus (see Table II) to provide the required shunt capacitance, no capacitors are necessary.

In some industrial plants, there may be individual motor loads, such as pump motors, tapped off a common overhead circuit at intervals along the line. For the best protection to all the motors the line should be shielded against direct strokes and special rotating machine protection provided at each motor. If shielding is not feasible, each motor should be provided with a series choke coil, special valve-type arresters and wave sloping capacitors at the motor terminals, and a standard line-type arrester ahead of the choke coil.

Rotating machines connected to the secondary windings of transformers whose primary windings are connected to overhead lines may not need rotating machine protection if the primary winding of the transformer is protected adequately. However, if large or important machines are involved, and are located close to the transformer, the risk of failure from surges passing through the transformer justifies the provision of special arresters and wave sloping capacitors at the machine terminals. Reference 5 includes a method of analyzing the need for such protection.

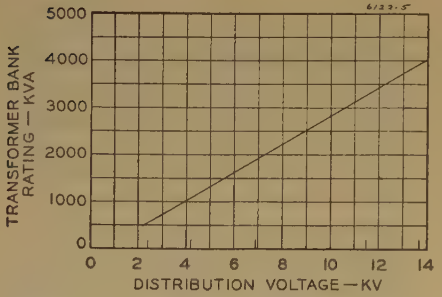

Figure 5. Approximate maximum sizes of transformer bank that can be grounded solidly without exceeding 2,500 amperes ground fault current

Protection Against Fault Damage

Even with good maintenance and surge protection there will be occasional faults, most of which will start as short circuits between one phase and ground. Such a fault may result in considerable damage by causing either severe arcing or high transient voltages, depending on whether the system neutral is grounded, and if so, how.[7–10]

Table II. Surge Capacitors or Equivalent Length of Cable Required for Rotating Machine Protection

Machine Voltage Class (Kv)	Capacitance Per Phase (Microfarad)	Cable Voltage Class (Kv)	Length of Cable (Thousands of Feet) 00	250,000 Cir. Mils	500,000 Cir. Mils	750,000 Cir. Mils
2.4	0.5	2–3	.35	.25	.20	.15
4.16	0.5	4–5	.45	.30	.25	.20
6.9	0.5	7–8	.50	.35	.25	.20
13.8 (grounded)	0.25	14–15	.30	.25	.20	.15
13.8 (ungrounded)	0.5	14–15	.60	.50	.40	.30

Neutral Point Available

If the neutral of the source of power to the distribution system, either generator or transformer bank, is grounded solidly, there will be no transient voltages of dangerous magnitude produced by a phase-to-ground fault. However, the ground current will be high, particularly if the grounded neutral machine is a steam-turbine-type generator supplying power directly to the distribution system. It will produce a noticeable dip in voltage and result in severe burning at the fault, unless the fault is cleared rapidly. It also may distort the windings of the grounded neutral generator which normally is designed to withstand short-circuit current no greater in magnitude than that resulting from a 3-phase short-circuit. The line-to-ground short-circuit current from a solidly grounded neutral generator usually will be greater than the 3-phase value.

If the source of distribution power is a transformer bank supplied from a high voltage line, grounding the neutral solidly will prevent dangerously high transient voltages resulting from ground faults. The magnitude of ground fault current will depend on the distribution voltage and the size of the bank. It will not cause excessive arc burning with moderate speed relaying if it does not exceed about 2,500 amperes. The approximate maximum sizes of transformer bank that can be grounded solidly without exceeding 2,500 amperes ground current are given in Figure 5. Higher ground currents are permissible if high speed relaying is provided.

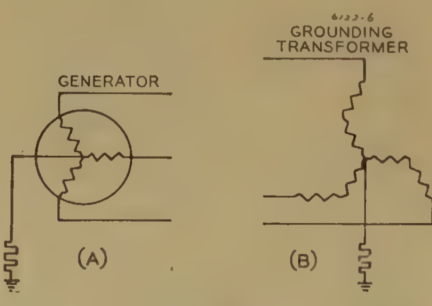

Figure 6. Low resistance neutral grounding

A—Generator neutral resistor
B—Grounding transformer with neutral resistor

If the neutral of the supply source is left ungrounded there will be no fault current when one phase becomes grounded except a few amperes of charging current through the circuit capacitance to ground. The extinction and restriking of this charging current may cause transient voltages between the unfaulted phases and ground high enough to destroy protective devices or to break down equipment insulation, thus causing one or both of the other phases to become grounded. A compromise between the high ground current and no transient voltages on the one hand and the high transient voltages and no ground current on the other hand is usually the best practical solution.

Neutral Impedance

The magnitude of ground fault current can be reduced by inserting impedance in the neutral ground circuit. However, care must be exercised in fixing the make-up of this impedance. If it is resistance only, there is a wide range in the value of resistance that may be used to eliminate the high transient voltages. It may be classed as either "low" or "high" resistance grounding (Figure 6) depending on the resultant magnitude of ground fault current. The lower limit of ground fault current for "low resistance" grounding is the minimum value that will assure dependable relaying with ground relays, about rated full load current with minimum generating capacity. The upper limit is dictated by the amount of energy that can be dissipated in the resistor during fault, both from the standpoint of practical resistor size and shock to the supply generators. A reasonable value to use for the upper limit is 150 per cent of the rated load current of the minimum generating capacity, or 2,500 amperes,

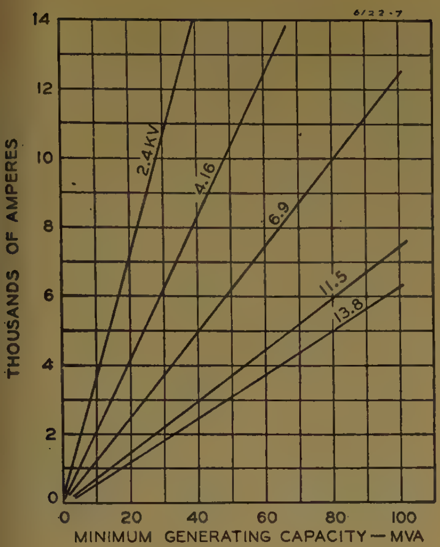

Figure 7. Approximate maximum values of ground fault current with neutral resistor grounding

whichever is the smaller. It usually will be desirable to approach this upper limit to assure prompt operation of ground relays under maximum fault resistance conditions. The curves of Figure 7 show values of ground fault current limited to 150 per cent of rated full load current for the voltages commonly used in industrial plant distribution systems. The minimum value of neutral resistance in ohms that should be used is equal to the line-to-neutral voltage divided by the value of fault current given by the curves.

From the standpoint of limiting the transient voltages, it is only necessary to use a neutral resistor of low enough resistance to dissipate energy equal in kilowatts to the charging kilovolt-amperes supplied the system at the time of a ground fault. This depends upon the maximum amount of circuit, particularly cable circuit, and the number and size of machines connected at any one time. With this value of neutral resistance, classed as "high" resistance, the current in the fault would be only about 50 per cent more than the charging current which would be but a few amperes at the most. The faulted circuit, therefore, could be left in service without danger from transient voltages or arc burning until convenient to remove it. In practice, an alarm actuated from a voltage relay connected to measure voltage between neutral and ground would be used to indicate when a grounded conductor existed. The circuit relaying, initially set up on the basis of ungrounded neutral service, would not be affected.

The effect of such a high resistance, neutral ground (Figure 8), can be provided more practically for 2,300 volts and higher by using a standard distribution transformer with a low voltage resistor connected across its secondary winding. The primary winding should have a voltage rating equal to the circuit voltage. It should be connected between the generator neutral and ground. The secondary winding may be 120, 240, or 480 volts, depending on the rating of the most applicable resistor which should be chosen to dissipate power during a line-to-ground fault equal in kilowatts to at least the maximum charging kilovolt-amperes drawn by the system during a ground fault. A voltage relay generally is connected across the resistor to serve as a fault detector.

When resistance grounding is used, the system neutral becomes displaced almost fully during a ground fault. This requires that lightning arresters connected to the system have a voltage rating not less than line-to-line voltage as is necessary with ungrounded neutral operation.

If the neutral impedance is reactance (Figure 9) there is a limit to how far the ground fault current can be reduced without causing high transient voltages. To be safe, this minimum value should not be less than about 50 per cent of the magnitude of the 3-phase fault current. The curves in Figure 10 show the minimum safe ground fault currents with reactance grounding for the different distribution voltages.

A combination reactor and resistor in parallel will permit high transient voltages unless the resistance is low compared to the reactance. Therefore, the two in parallel offer very little, if any, advantage over a resistor only.

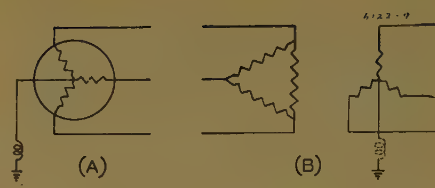

Figure 9. Reactance grounding

A—Generator neutral reactor
B—Transformer with or without neutral reactor

Any neutral ground scheme that permits enough ground fault current to cause arc damage requires ground relays. If the neutral is grounded solidly, the ground fault current would in most cases be high enough to operate phase relays. However, there may be occasional high resistance ground faults that would not draw enough current to operate phase relays in which case considerable damage from burning might result if ground relays were not provided.

Neutral Point Not Available

Many industrial plant distribution systems do not have a neutral point available for grounding, particularly those that operate at 2,300 or 6,900 volts, delta. As previously pointed out, such systems are conducive to high transient voltages during ground faults. Experience has shown that a failure of a machine winding or cable circuit to ground often results in the simultaneous failure of other machines or cables. The remedy is to supply one or more supplementary neutral grounds to assure that all feeder sources are grounded adequately at all times.

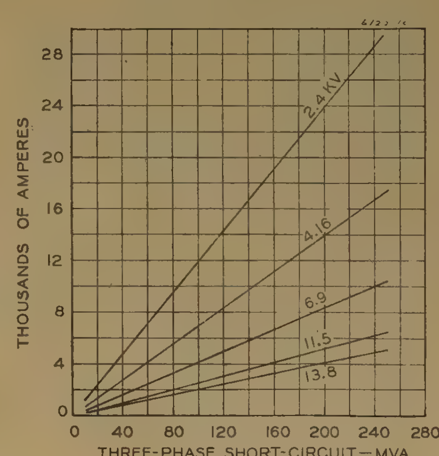

Figure 10. Approximate minimum values of ground fault current with reactance grounding

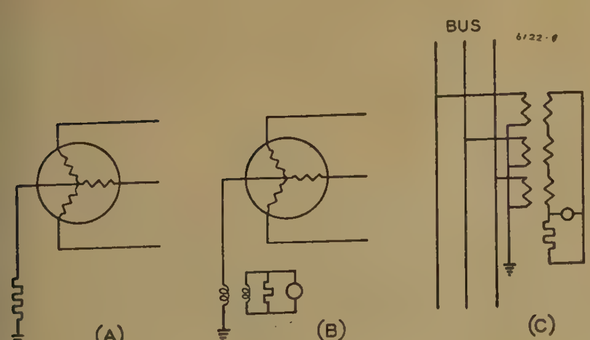

Figure 8. High resistance neutral grounding

A—Generator neutral resistor
B—Distribution transformer in generator neutral with secondary resistor
C—Bank of distribution transformers with secondary resistor

If the distribution system draws only a few amperes charging current and is relatively simple in make-up so that a grounded conductor can be located without too much difficulty, a high resistance neutral ground may be established with a small bank of distribution transformers and secondary resistor as shown in Figure 8*C*. The star-connected primary windings should be connected to the distribution bus with neutral grounded. The resistor should be connected inside the delta of the low voltage winding. The resistance should be chosen to dissipate energy equal in kilowatts to at least the charging kilovolt-amperes of the system during a ground fault. This scheme of grounding is similar to the distribution transformer–secondary resistance scheme described previously, in that it serves only to limit the transient voltages and to indicate a ground fault. The fault current is too low to cause arc burning so that ground relays are not necessary.

The other method of providing a supplemental neutral ground is to use a grounding transformer, either zig-zag or star delta connected with the neutral grounded. If grounded solidly it is classified as reactance grounding and carries the same limitation as described previously, that is, that the ground current should not be less than about 50 per cent of the 3-phase fault current. If grounded through a resistor (Figure 6*B*) the ground current can be reduced to the minimum permissible for satisfactory ground relaying. If a "low resistance" neutral resistor is used, the reactance of the grounding transformer should be such that if the transformer were grounded solidly, the ground fault current would not be less than about 50 per cent of the 3-phase value. Relays that will clear ground faults promptly are necessary to prevent excessive arc burning.

Preferred Method of Neutral Grounding

The essential requirement of a neutral ground is that it will suppress high transient voltages during ground faults. The scheme best suited to any particular industrial plant depends on a number of factors, such as the surge protection required, the magnitude of the power source, whether a neutral point is available or not, the number and complexity of circuits involved, and the type of relaying that is or can be provided. From the standpoint of surge protection, either solid neutral or reactance grounding that limits the ground fault current to not less than 60 per cent of the 3-phase fault current is preferable, because either one makes it possible to apply reduced rating (80 per cent) arresters. If the insulation is questionable, this may be the deciding factor. However, either of these types of grounding produces high ground fault currents if the distribution system is supplied at generator voltage and the generating capacity is large. When those conditions exist, resistance grounding generally will be suited better. The distribution transformer–secondary resistor scheme often works out advantageously where the distribution system is not too complex and where it is not feasible to add ground relays.

Conclusions

There are methods available for surge protecting almost any combination of circuits and apparatus that may be encountered in industrial plants. Consequential damages due to high transient voltages resulting from unavoidable ground faults can be prevented by proper neutral grounding. Experience has shown that, where these precautions are taken, trouble from electrical failures has been reduced to a minimum.

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Fundamental Properties of the Vacuum Switch

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Synopsis: The main results of a study of the various phenomena which take place during the interruption of currents in high vacuum are given. The experimental results obtained lead to a theory of the basic mechanism affecting switching in high vacuum. With a proper design of the switch, a constant working pressure can be obtained which represents an equilibrium between gases and vapors evolving from the electrodes and a pumping action based on the sputtering of cathode material. A detailed study is made of the factors determining the loss of electrode material because of switching. The most essential design features of a vacuum switch are discussed briefly.

SWITCHING is one of the most frequently performed operations in electrical circuits. Even though switching plays a very important role, the equipment performing this operation, the switches and circuit breakers, have by no means reached the last degree of perfection.

No satisfactory d-c breaker has been developed yet for the interruption of medium or heavy loads at high voltage.

The interruption of alternating current is made much easier by the periodical occurrence of the current zero which represents a natural means of deionization. The difficulty here lies in the fact that severe requirements are made with respect to the time of interruption.

System stability studies indicate that faults on transmission lines, particularly of large interconnected systems, should be cleared in a minimum of time. This is the main reason why the increase in load of the past 15 years gradually has placed stronger requirements on the opening time of breakers. Whereas in 1930 a 10- to 15-cycle breaker was found to be satisfactory, today the tendency is to clean faults in three cycles. It is to be expected that in the near future still more severe specifications will have to be made on the interruption time.

In some cases no restrikes after the first current zero are permissible, as for example in the interruption of charging current. Restrikes would result in high overvoltages, exceeding the normal line voltage several times.[1,2] This is particularly dangerous on power lines where sections of cable are incorporated.

With the present designs of circuit breakers, it is very difficult to meet requirements of the type mentioned above. The main difficulty encountered by the present methods of arc extinction lies in the small dielectric strength of the interrupting medium. The rate of recovery of dielectric strength is small, even when large opening speeds are used.

It was recognized early that the use of high vacuum as an interrupting medium is expected to have two features which are of great importance for the rapid interruption of currents:

1. The dielectric strength of high vacuum[3,4,5] is greater by far than that of any other interrupting medium in use.

2. Once an arc has been formed in a high vacuum, a fast radial diffusion of the current carrying particles takes place and provides a most ideal means of deionization.

A great deal of research and development has been done on vacuum switches. In 1921 the first small vacuum switch was built commercially by the Swedish Birka Company.[6] From 1923 to 1926, Sorensen and Mendenhall[7] performed extensive tests with larger vacuum switches at the California Institute of Technology. Around 1932 interesting investigations were carried out by Seitz[8] and Jubitz[9] in Germany.

As a result of these efforts, excellent small current capacity vacuum switches are commercially available and are finding a steadily increasing field of application.

Today, however, the vacuum switch is not able to compete in power circuits. No explanation for this fact can be found in the literature. There is also no theory given explaining the various processes affecting the interruption in high vacuum.

The aim of the investigation described in this paper was to clarify some of the important and basic points involved in this problem.

Experimental Setup

The investigation was carried out entirely on direct current. The interruption of alternating current only would have complicated the analysis greatly without giving any new information about the basic mechanism of the interruption.

The two switches shown in Figures 1 and 2 were built for this study. The switch of Figure 1 is completely glass-sealed, allowing studies to be made at very low pressures of 10^{-6} millimeter of mercury or less. This switch is actuated magnetically by means of a nickel plunger and a trip coil. The second switch, shown in Figure 2, was developed after it had been established that pressures below 10^{-4} millimeters of mercury were not required for a successful operation. Two main improvements were incorporated in this switch. First, it provides a simple and quick way of opening and resealing. Picein was used as a sealing wax around the steel base and the ground bottom of the glass jar.

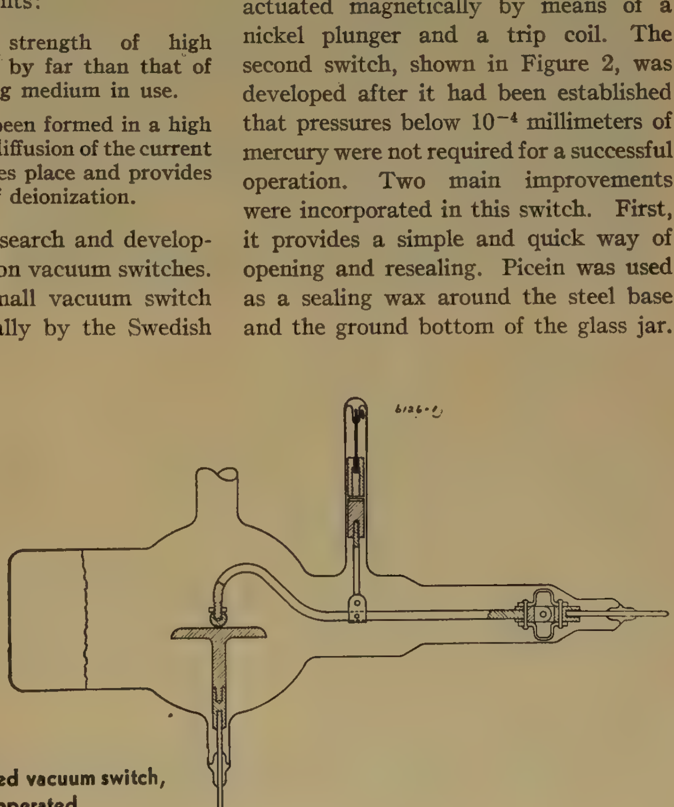

Figure 1. Glass-sealed vacuum switch, magnetically operated

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The second improvement consisted of an accurate and easily changeable speed of separation of the electrodes, which is a significant variable. A spring located in the socket of the lower electrode provided the desired contact pressure. When the switch was tripped, the lower electrode followed the parting upper electrode until a stop in the socket was reached. At this point the lower electrode was arrested and the electrodes separated.

The switch is actuated by an outside mechanism shown in Figure 3. The motion is transmitted to the switch by means of a metal bellows located at the top of the switch.

The disk D of Figure 3 is rotated at synchronous speed. The opening spring Sp is expanded by means of a pin A mounted eccentrically on the disk B. A short time before the maximum stretching of the spring is reached, the cam switch SW closes and energizes the coil and the magnet M. This locks the tripping toggle T by means of the latch L. The tripping coil TC releases the tripping toggle. This results in a rapid opening of the switch. The trip coil can be energized by closing the switch S or through the cam switch CS. This cam switch permits the automatic operation of the vacuum switch. This is important because thousands of operations are required to determine accurately the average loss of electrode material per interruption, which is very small.

The vacuum system used is illustrated by Figure 4. The vacuum is created by a forepump F and a 2-jet mercury diffusion pump J. Another forepump M was used to reduce the time required for the complete evacuation of the vessel. The use of this pump for the first part of the evacuation also helped greatly to keep the two liquid air traps clear of ice formation.

The pressure in the vessel was measured with a cold cathode ionization gauge I of the Penning[10] type. The McLeod gauge G was used for checking purposes only.

The speed of separation of the electrodes of the spring operated switch was measured with a mechanical speedograph.

Another method had to be devised for the measurement of the opening speed of the magnetically operated switch. There the force available from the magnetic plunger was too small for the use of a speedograph. An optical method illustrated in Figure 5 was used instead.

A narrow vertical beam of light is obtained by means of the condenser lens L_1 and the slit S. In the closed position of the switch, this parallel beam of light is obstructed completely by the electrodes. During the separation, a beam of light which is proportional to the width of the gap reaches the lens L_2 and is focussed on the phototube. The resulting phototube current then is amplified in order to give a satisfactory deflection on the screen of a cathode ray oscillograph.

A curve can thus be obtained on the screen, giving the separation of the electrodes as a function of time. Such curves used in conjunction with recovery voltage oscillograms allow one to determine the electrical field strengths present during interruption.

The Mechanism of Interruption

In what follows, a detailed analysis of the mechanism of interruption is given. This analysis is based on the various observations made by means of different methods, such as cathode ray oscillograms, fast movie films, microscopic studies of the electrode surface, and so forth.

At the beginning of the separation of the electrodes, the contact area is decreased rapidly. This results in a fast increase in current density, leading eventually to melting and evaporation of the electrode material. A vapor arc thus is created which rapidly blows itself out because of the rapid radial gaseous diffusion active in the evacuated space. The loss of current carriers because of this radial diffusion is large enough to disrupt quickly the self-sustaining discharge.

The average velocity of diffusion is given by[11] $C = \sqrt{4kT/m}$, where k = Boltzmann constant, T = absolute temperature, and m = mass of a gas molecule. At a temperature of a few thousand degrees Kelvin, the diffusion velocity is of the order of 10^5 centimeters per second, which is about 1,000 times greater than the speed of separation of the electrodes.

As soon as this first arc is blown out, a recovery voltage builds up across the gap, which is by now only about 10^{-}

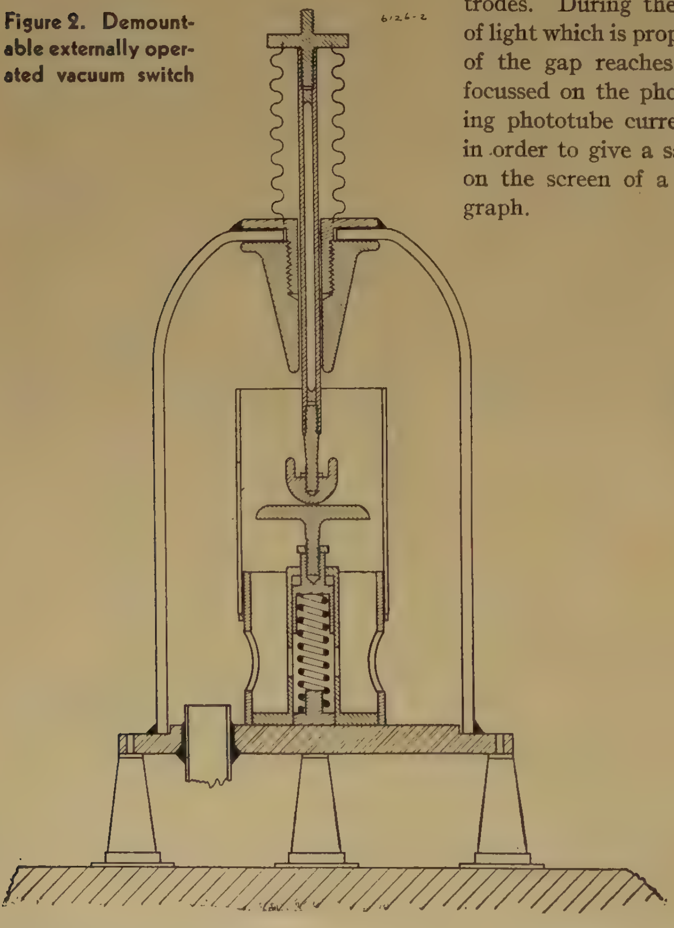


Figure 2. Demountable externally operated vacuum switch

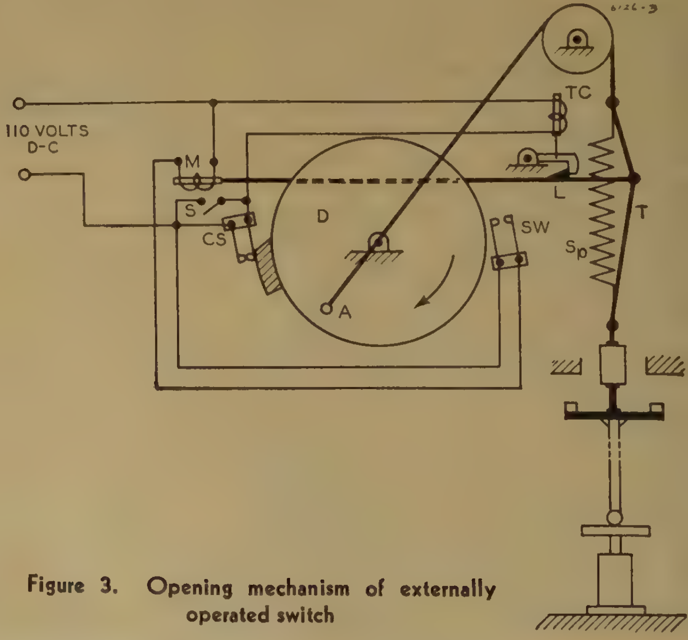


Figure 3. Opening mechanism of externally operated switch

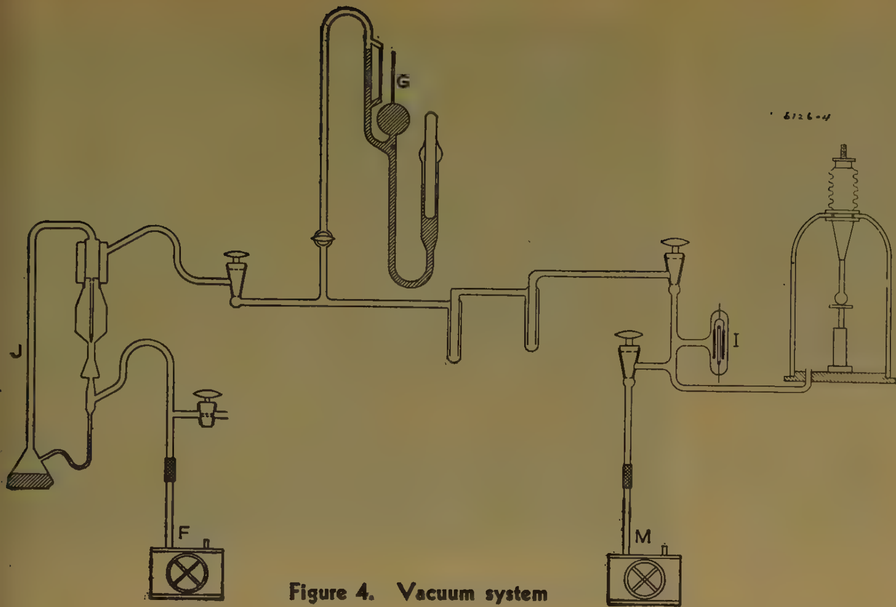


Figure 4. Vacuum system

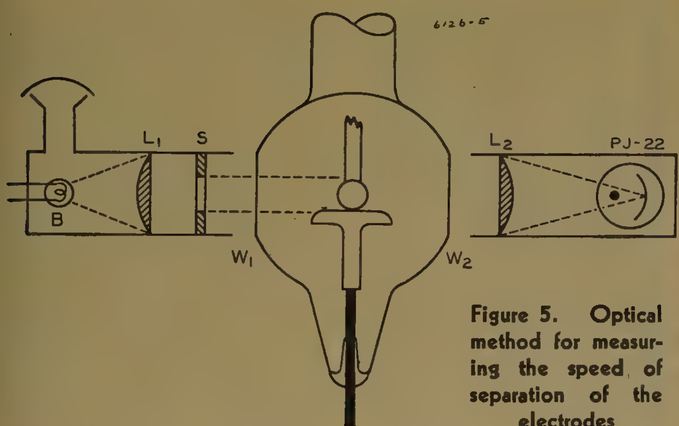


Figure 5. Optical method for measuring the speed of separation of the electrodes

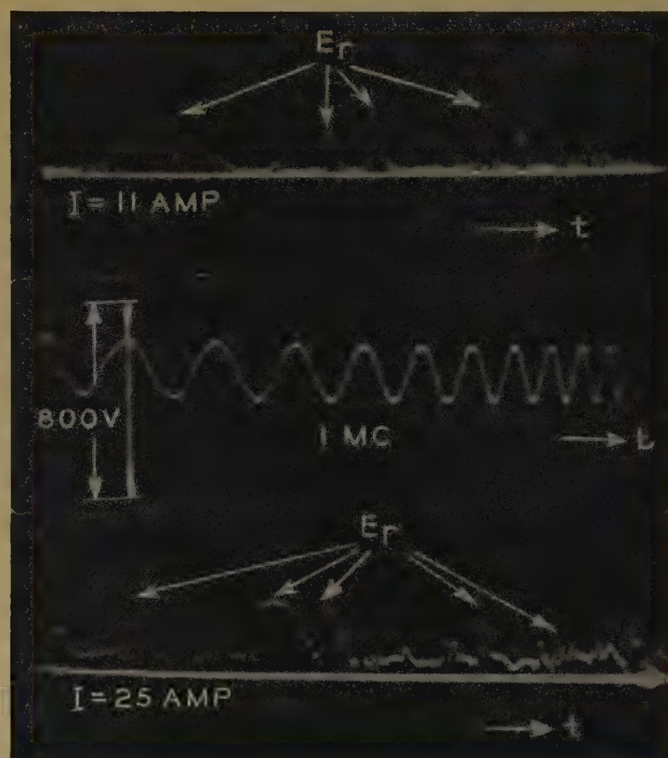


Figure 6. Voltage oscillograms with one megacycle timing wave—copper electrodes

centimeters wide. Under this condition a relatively small recovery voltage creates a field strength of sufficient magnitude to break down the gap in the form of a special kind of spark arc mechanism to be discussed later. The initiating spark preferably will take place along a path where a small density of gas and vapors is left, that is, about one millimeter away from the former point of contact, owing to the diffusion in the microsecond of the first arc. The arc following the initiating spark is blown out again, and another recovery voltage E_r leading to a new breakdown builds up across the gap. Recovery voltages E_r of this type are well resolved in the oscillograms of Figure 6. It can be seen that the mechanism consists of a series of arc bursts of microsecond duration, interspersed with periods of extinction of even shorter duration. It is significant that the random character of this process is found by experiment not to be affected by the circuit constants.

For the case of a plate-cathode and a spherical anode, fast movie films show that the spark-arc mechanism moves away from the center of contact along a single radial line. This situation is represented schematically in Figure 7. It can be seen that the radial diffusion of the gas and vapors originated during each single arcing episode A_1, A_2, A_3... creates ring-shaped zones of greatest gas vapor and ion density. The successive breakdowns B_1, B_2, B_3...following the recovery voltages E_1, E_2, E_3...take place in the regions where the ring-shaped zones overlap, that is, along a single radial line.

In the case of large currents, spark-arc mechanisms will occur along many separate radial lines. The separate arcs occurring simultaneously at a given time all will be located on a concentric circle around the former point of contact. This form of discharge along an arc of a circle is found to take place for currents exceeding about 30 amperes.

The interruption is completed when such a series of arcing episodes ceases. This happens once the path along the greatest gas density is too long, that is,

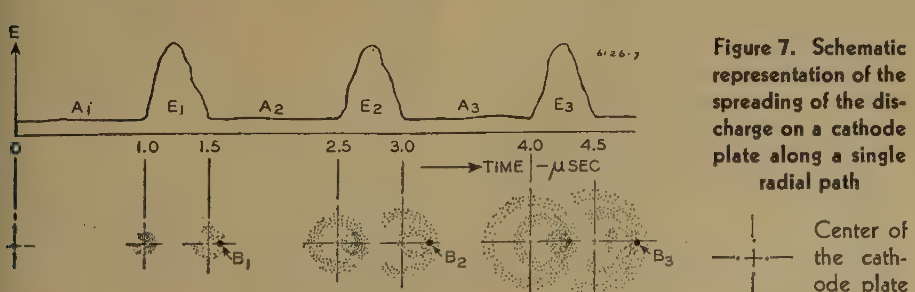


Figure 7. Schematic representation of the spreading of the discharge on a cathode plate along a single radial path

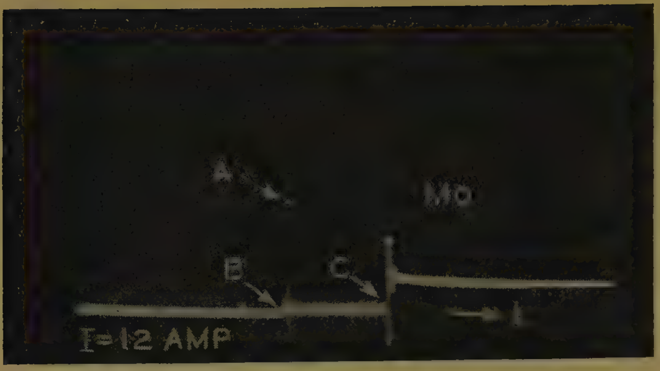


Figure 8. Voltage oscillogram—molybdenum electrodes

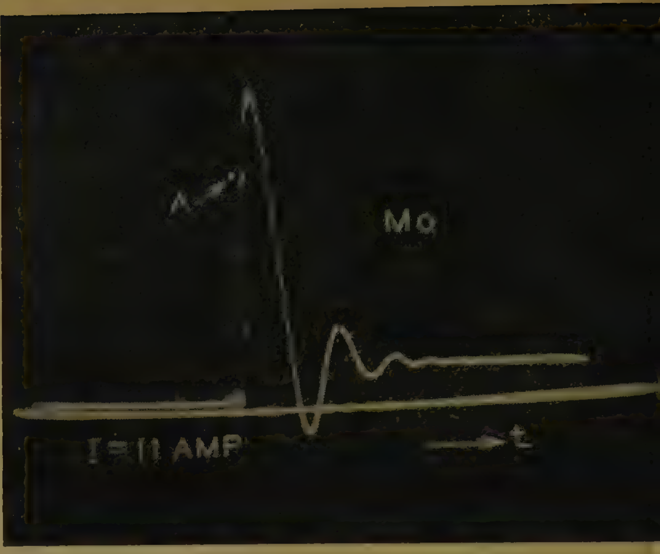


Figure 9. Voltage oscillogram—molybdenum electrodes

when its dielectric strength prevents a further breakdown. At the end of such a series of arcing episodes, the electrodes are still close together. If the circuit conditions create a high recovery voltage, the gap will break down near the center by field emission. This usually starts another succession of radial spark-arcs. Fast movie films were able to give evidence of repeated breakdowns near the center of the gap. The field strengths corresponding to such breakdowns exceed 10^6 volts per centimeter. It is therefore clear that this type of breakdown is initiated by field emission.[12]

Figure 8 shows a typical voltage oscillogram where a high recovery voltage occurred at point A, 3×10^{-5} second after the start of the interruption at point B. The total time of interruption (B to C) amounted to 3.2×10^{-4} second. The recovery voltage at point A was 552 volts, corresponding to a field strength of 2.76×10^6 volts per centimeter. The succeeding large recovery voltage which is clearly resolved occurred at point C and amounts to 1,380 volts, giving a field strength of 1.26×10^6 volts per centimeter. Large recovery voltages are always present at the end of the interruption of an inductive d-c circuit. The resulting field strengths are, however, no longer able to restrike. Figure 9 shows an interesting case where a field emission breakdown occurred at point A which was not able to reignite a new series of arcing episodes.

The fact that the mechanism of interruption is affected by field emission explains the observed great variations of the time of interruption. The determination of the statistical average time of interruption requires, therefore, the data of a large number of interruptions.

The study of electrode surfaces with the microscope revealed another interesting phase of the mechanism of interruption. Figure 10 shows the many microscopic craters of about 10^{-3} centimeter diameter which have been created on a tungsten cathode. The cathode also shows cracking and peeling of the surface which was caused by the heat resulting from the intense positive ion bombardment. The anode shows many fused spots; the rest of the surface is covered with sputtered material from the cathode. Carbon electrodes show even more distinctly the difference between cathode and anode surfaces. Figure 11 shows how the cathode surface is broken up to a great depth by the combined action of the positive ion bombardment and the explosive action of the suddenly liberated gas.

The anode has a smooth surface covered with sputtered carbon from the cathode. The microscopic craters observed on all cathodes reveal the presence of a focusing action. The conditions present during the periods of recovery voltage E_1, E_2, $E_3 \ldots$ (see Figure 7) as well as during cold emission restrikes are leading to the conclusion that a magnetic self-focusing action of the Bennett[13] type takes place. This magnetic self-focusing concentrates the positive ion bombardment on small areas on the cathode, resulting in local fusion and evaporation. This creates microscopic craters of the type indicated by arrows in Figure 10. The self-focusing action lasts only about one microsecond. After this interval of time, the conducting streams are no longer self-focusing because the density of matter near the cathode leads to frequent collisions between current carriers and neutral atoms. The conditions for self-focusing hold, however, only as long as the energy loss caused by such collisions is negligible. As soon as the focusing action ceases, a transient arc of short duration is established. This arc is quickly deionized by the fast radial gaseous diffusion. This then is followe[d] again by a series of arcing episodes a[s] described earlier.

Variations of Pressure Produced by Switching

Contrary to previous statements, i[t] has been found that the vacuum switc[h] normally operates at a vessel-pressur[e] of 10^{-4} millimeter of mercury or less provided that the electrodes are wel[l] outgassed.

Throughout this investigation, stock-material was used for the electrodes. The outgassing or conditioning of suc[h] electrodes was achieved in interruptin[g] frequently with gradually increasin[g] currents. The heat created by radiatio[n] and particularly by the positive io[n] bombardment of the cathode was suffi-cient to free the electrodes from adsorbe[d] gases. A satisfactory conditioning o[f] the electrodes required about 10[0] operations. All interruptions have t[o] be started at a vacuum in the vessel o[f] 10^{-4} millimeter of mercury or better. The use of vacuum-melted material of course, would shorten considerabl[y] the time required for the conditioning.

Once the electrodes are conditione[d] properly, it was found to be of no ad-vantage to operate the switch at pressure[s] below 10^{-4} millimeter of mercury. Highe[r] pressures result in a considerably longe[r] time of interruption, as will be show[n] later.

An important property of the vacuu[m] switch is the fact that the pressure i[n] the vessel does not necessarily increas[e] as a result of switching. With the vesse[l] disconnected from the pumping system, [a] definite constant working pressure estab[...]

lishes itself in the vessel as a result of frequent switching. In the course of the experiments, the vacuum switches often were disconnected from the pump for hours, performing thousands of identical operations, without resulting in a change of the vessel pressure.

The working pressure was found to vary with the vessel diameter, the current, and the electrode material. The working pressure even can become appreciably lower than the pressure which is present in the vessel before the interruptions of currents. This behavior indicates that a strong pumping action is initiated by the interruption of currents. The observed constant working pressure thus represents an equilibrium between this pumping action and the gases and vapors which necessarily evolve from the electrodes. Real and virtual leaks, of course, also will affect this equilibrium. The pumping action is based on two distinct processes, both of which result from the sputtering of the cathode material. One process acts like a diffusion pump, whereas the other is based on the getter action which some materials like molybdenum display very strongly. The mechanism of the diffusion pump like process is as follows: The sputtered cathode material diffuses at high speed to the walls. Gas molecules present between the discharge region and the walls then are caught by the diffusing metal atoms and are plastered on the walls. These gas molecules are bound firmly to the walls and can be freed only when the walls are heated above 300 degrees centigrade, as is well known from the case of glow discharge tubes. Pumping in the manner described will occur when the distance from an active spot on the cathode to the nearest section of the wall is of the order of the mean free path of the gas in that region. This mean free path is a measure for the collision probability between the gas molecules and the sputtered particles. Under the conditions of the experiment, it always was found that the mean free path corresponding to the pressure in the vessel (measured with the ionization gauge) was several times larger than the average distance S of the active spots to the wall. This can be explained by the fact that as a rule the evaporating metal atoms greatly outnumber the gas molecules in the volume exposed to the blast of sputtered material. For example, during an interruption of 25 amperes with copper electrodes, the evaporated copper atoms outnumbered the gas molecules by a factor of fifty. The fact that the mean free path is larger than the distance S means that only a fraction of the metal atoms succeed in plastering gas molecules to the walls. The actually resulting reduction of gas molecules, partially balanced by the evolution of gases from the electrodes, thus will maintain a lower gauge pressure than that corresponding to a mean free path equal to the distance S.

It was found that the working pressure p, as measured by the gauge, varies inversely with the baffle or vessel diameter d, as $p_1/p_2=d_2/d_1$.

This relation holds well for materials like copper and aluminum which do not display a strong getter action. Thus the judicious use of baffles may be of value in the switch design.

The current does not affect greatly the working pressure since it equally affects two opposing processes: the liberation of gas from the electrodes and the amount of sputtered material. The electrode material has a similar effect on the working pressure in affecting the evolution of gas as well as the sputtering. Materials which sputter very little (molybdenum, tungsten) require larger currents in order to display a strong pumping action.

The getter action also can contribute to an appreciable reduction in pressure. The fine film of sputtered material which collects on the walls is very efficient in absorbing gas molecules. Molybdenum displays a particularly strong getter action which was observed to last with diminishing intensity for 30 minutes or longer.

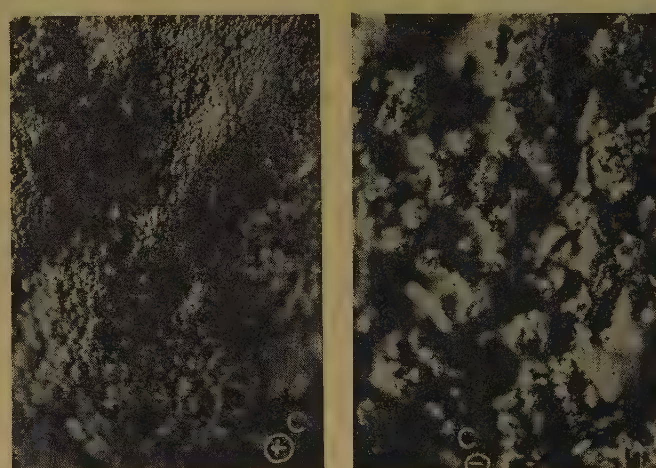
The reduction in pressure resulting from the two combined pumping actions can be considerable and is in part responsible for the fact that a current can be interrupted repeatedly in vacuum.

Anode Cathode

Figure 10 (above). Molybdenum electrodes (× 90)

Figure 11 (below). Carbon electrodes (× 70)

Anode Cathode



Factors Determining the Loss of Electrode Material

From the knowledge of the mechanism of interruption, it follows that the loss of electrode material caused by the interruption of currents will depend on the mass of material sputtered or evaporated at each striking of the arc and on the number of individual arcing episodes occurring during the time required for the interruption of the circuit. The mass of material sputtered at each striking of the arc is probably proportional to the energy dissipated in such an arc. The amount of mass sputtered per watt-second, however, will depend particularly on various physical constants of the electrode material and therefore will differ a great deal among various ma-

terials. For a given material the losses must thus depend on the over-all time to terminate the interruption and on the currents and potentials involved.

From the mechanism of interruption described earlier, it follows that the primary loss of material comes from the cathode. According to observations of the electrode surface, it is to be expected that the anode gains in mass by the sputtering of cathode material onto it.

Unfortunately, too little is known today about the mechanism of sputtering to enable one to make a thorough quantitative analysis of the various factors influencing the loss of electrode material. Two distinctly different types of sputtering have been observed in the course of the experiments. One is the atomic sputtering resulting from the evaporation of electrode material. The other type of sputtering is caused by occluded gases in the electrode material. The positive ion bombardment produces a fast local heating of the cathode surface which results in an explosive gas evolution and in the blasting out of chunks of material from the cathode. There has been visible evidence of such incandescent tiny chunks streaking from the arc region toward the walls and often rebounding from them. Carbon contains great amounts of absorbed gases and therefore displays with great intensity this second type of sputtering. The diffusion pump like action, described in an earlier part of this paper, is caused entirely by the atomic type of sputtering.

In what follows, the experimental results of the investigation on electrode material losses are given.

Distribution of Sputtered Material

All switching experiments resulted in a gain in mass of the anode and in a loss in mass of the cathode. The material sputtered from the cathode deposits on the anode, on the walls of the switch, and is reflected partially to the cathode. The loss of anode material which is known to occur from electron streams is overcompensated largely by the cathode material collecting on the anode surface.

The distribution of the sputtered material is determined chiefly by the design of the switch and particularly by the size and the shape of the electrodes. However, it also is affected slightly by the current, the recovery voltage, and the speed of separation of the electrodes.

Influence of the Gas Pressure

During this investigation, the pressure in the vessel was brought to a definite value before each interruption. Insignificant changes in losses were observed in interrupting at pressures ranging from 10^{-6} to 10^{-4} millimeter of mercury. At pressures above 10^{-4} millimeter of mercury, the losses increase rapidly because of a lengthening of the time of interruption.

If the pressure becomes too high, a permanent arc was able to maintain itself and switching action ceased. The lengthening of the time of interruption with increasing pressure obviously is caused by a corresponding decrease of the velocity of the radial gaseous diffusion.

Influence of the Shape of the Electrodes

The size of the anode was found to have no noticeable effect within the range investigated, that is, from 0.5 centimeter to 50 centimeters diameter. The curvature of the cathode has a pronounced effect, in increasing the loss and the time of interruption for a smaller radius of curvature. With a small radius of curvature, the radial travel of successive arcing episodes leading to extinction is no longer possible as it was in the case of a plate cathode. The restrikes, therefore, will be confined to a limited area around the former point of contact until the separation of the electrodes is too great to restrike the arc. This localization of the interruption process results in more continuous arcing episodes with a deeper localized heating of the area. This yields more gas evolution and evaporation and consequently heavier instantaneous arcing-currents result.

Influence of the Vessel Diameter

It was found that the loss of cathode material was slightly higher for smaller vessel- or baffle-diameters. The effect is very small and can be explained by the increase of the working pressure, resulting from the smaller diameter.

Figure 12. Average cathode loss versus average time of interruption of a d-c circuit

Cathode: 2-inch diameter plate
Anode: 1-inch diameter sphere
Opening speed 75 centimeters per second

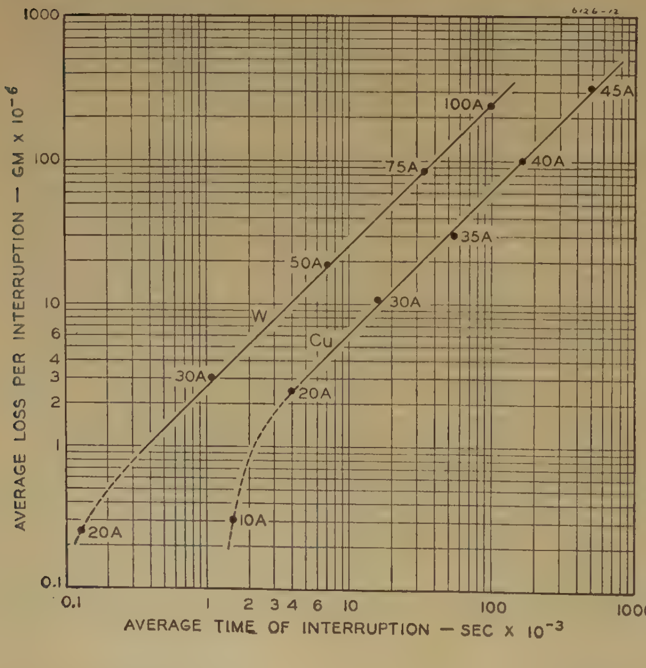
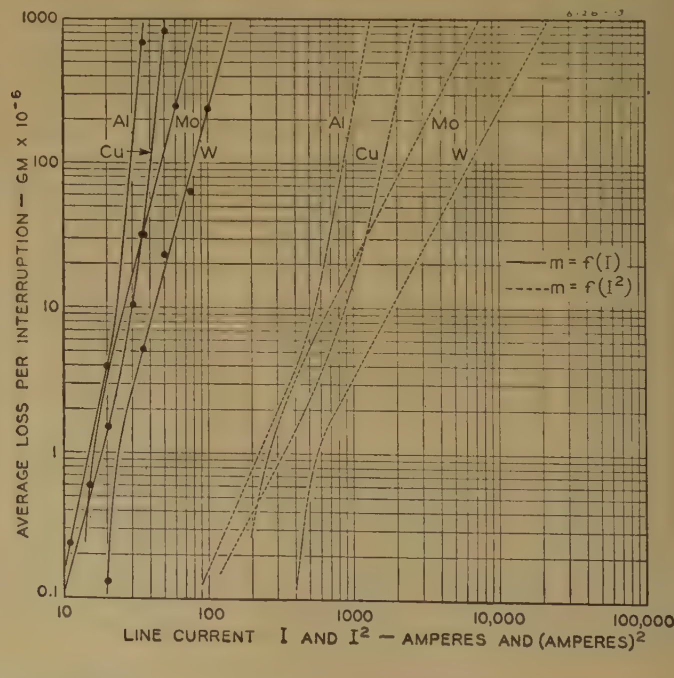


Figure 13. Average cathode loss versus line current I and I^2 of a d-c circuit

Cathode: 2-inch diameter plate
Anode: 1-inch diameter sphere
Opening speed 75 centimeters per second



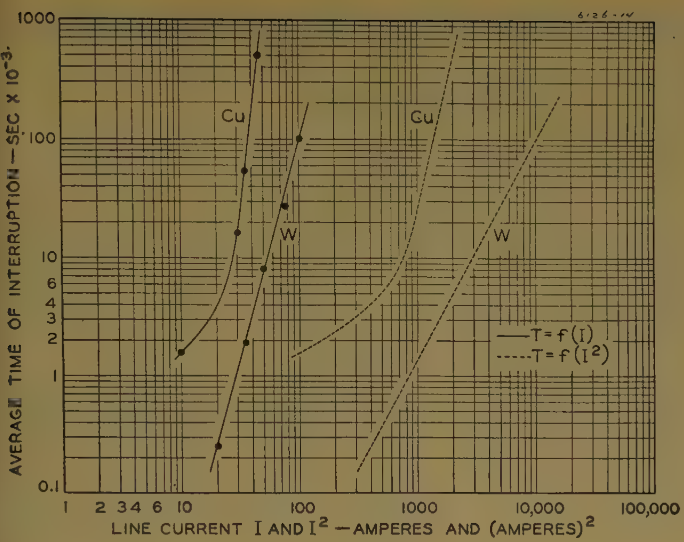


Figure 14. Average time of interruption versus line current I and I^2 of a d-c circuit

Cathode: 2-inch diameter plate
Anode: 1-inch diameter sphere
Opening speed 75 centimeters per second

The opening speed was in each case 75 centimeters per second. Referring to Figure 13, it is seen that the refractory metals give the smallest losses in the high current region. At lesser currents the different metals behave in a different fashion, the more refractory metals indicating a greater slope and the low melting point metals a smaller slope. It appears that two mechanisms may be at work in different current ranges. The difference in losses between the higher and lower melting point metals indicate why good conducting metals like silver and copper often are used in low current capacity vacuum switches.

The line current has a two-fold effect on the losses. Considering the energy equation of the circuit, it follows that the current determines essentially the sum of $1/2CE_r^2+A$. It affects the recovery voltage E_r as well as the energy loss A. The latter term is particularly affected by the amount of evolved gases and vapors, which depend on the current magnitude and which greatly affect the time of interruption. It is obvious that well outgassed electrode materials, giving off only small amounts of gas, will result in small losses. It therefore is expected that vacuum-melted materials would give considerably smaller losses than can be expected from Figure 12 and Figure 13.

The surface conditioning as used throughout the present investigation is not sufficient for currents exceeding 150 to 200 amperes where craters are dug to a depth which is beyond the region outgassed by previous conditioning. This results in excessive gas formation, which would be dangerous, particularly for the case of sealed-off switches.

Influence of the Line Voltage

The line voltage had no effect on the losses within the range investigated, that is, from 50 to 500 volts direct current. This is also in accordance with the findings of Sorensen and Jubitz. The line voltage is not of great significance in any such phenomena inasmuch as the recovery voltage is usually much higher. The recovery voltage is a much more significant variable and acts in two ways. One of these is to increase the chance of restriking and to increase the number of arcing episodes, thus lengthening the time of interruption. The other is directly to affect the sputtering. In the operation of glow discharge tubes, it is well known that sputtering increases materially once one enters the region of abnormal cathode fall, where the voltage increases with increasing current.

Influence of the Speed of Separation of the Electrodes

A few simplified considerations have to be introduced in order to understand the effect which the opening speed has on the loss of electrode material.

The energy equation[8] of a circuit having capacity C and inductance I is as follows:

$$1/2LI^2=1/2CE_r^2+A$$

where I=line current, E_r=recovery voltage, and A=loss of energy in the circuit and in the switch because of heat and radiation. The term on the left side of the equation represents the magnetic energy stored in the system at the time of interruption. It is a constant for a given current. Referring to Figure 12, it is seen that except for small currents the loss of cathode material is directly proportional to the time of interruption T. Therefore the loss of energy A in the circuit is also closely proportional to T. For a given current and a given set of electrodes, a definite separation d of the electrodes is necessary for the interruption of the circuit. Hence $d=vT$, where v is the speed of separation of the electrodes.

If d would remain constant, then T would be inversely proportional to v. The experiments, however, have shown that the mass-losses of the cathode are always less than proportional to $1/v$. The reason for this is that an increase in the opening speed v tends to reduce A in making T shorter. Since the right hand side of the above equation is constant, it follows that the recovery voltage E_r has to increase. This actually has been observed. A greater separation d_1 of the electrodes therefore is required for the interruption of the circuit.

Since d is not a constant, it follows that an increase in velocity will always result in a decrease in mass loss which is less than proportional to $1/v$.

Influence of Line Current and Electrode Material

Figure 13 shows the average cathode material loss for various electrodes and different values of current flowing in the line prior to interruption.

In the same figure the broken-line curves show the relationship between cathode losses and I^2. This is a measure of the magnetic energy $1/2LI^2$ stored in the circuit before the time of interruption. Figure 14 shows the relation between the average time of interruption T and I as well as I^2.

The electrodes used for all the data given in Figures 12, 13, and 14 consisted of a 2-inch-diameter plate cathode and of a 1-inch-diameter spherical anode.

Essential Design Features

The mechanism of interruption, the behavior of the pressure, and the character of the losses of electrode material indicate that improvements can be made in the design of vacuum switches for moderate- and high-interruption duties.

The fact that a very high vacuum is not essential for the proper operation makes a practical design of a vacuum switch possible.

The vacuum vessel can be made of glass or metal, preferably stainless steel. The seals can be of the enamel type; welded, brazed, or soft soldered joints are also satisfactory. Glass-to-metal joints have to be ground and can be made vacuum-tight by means of a low-vapor-pressure wax, like picein. Greases should be avoided because they dissociate easily

under the intense radiation of the discharge.

The switch requires, for conditioning, a relatively low capacity diffusion pump backed by a medium-sized fore pump. The cooling of the traps can be achieved satisfactorily by means of a mixture of dry ice and alcohol.

It is expected that the judicious choice of seals and materials will make a permanently sealed-off vacuum switch possible. In this case it will be of even greater importance to use the correct vessel or baffle diameter by means of which full use of the pumping action can be made. Metal or glass baffles are also useful in shielding the insulating parts of the switch from sputtered material. A thorough outbaking will be required for a sealed-off switch.

The vacuum switch is ideally suited for the interruption of high voltage circuits. It is necessary, however, to reduce the evolution of gases and vapors as much as possible. Therefore, it is necessary to limit the current per contact and to choose electrode materials with small gas and vapor evolution. For larger currents, refractory metals are best suited.

The electrodes have to be conditioned in order to reduce the gas evolution. The conditioning by positive ion bombardment, as described earlier, limits the current to from 100 to 200 amperes per single contact. For larger currents, vacuum processed metals have to be used.

Exposure to air during the insertion (of new contacts, for instance) tends to contaminate the surface of such materials and therefore indicates the necessity of a short conditioning procedure preceding the operation of the switch after each shutdown period.

The interruption of higher currents and voltages also can be accomplished by multiple contacts. Such contacts will have to break with a fair degree of simultaneity.

All current carrying parts exposed to the vacuum should have large cross sections in order not to be heated appreciably by the normal line current or by short circuit currents of short duration. Too high a temperature of these elements would result in an undesirable evolution of gases from their surfaces.

The small distance of separation of the electrodes required for the interruption allows tripping the switch with great accuracy. With alternating current it is thus possible to interrupt close to current zero. In this way only small currents and recovery voltages have to be handled by the switch, resulting in a small loss of cathode material.

The loss of material and the time of interruption can be influenced further by the speed of separation of the electrodes. A speed of one meter per second or more should be used at the instant of separation.

The radius of curvature of the cathode should be large. The shape of the anode is not essential. For the interruption of alternating current the same contact arrangement as used for direct current is applicable provided the switch is always tripped at the same polarity. Both electrodes have to have a large radius of curvature and have to be symmetrical if the switch interrupts on either polarity.

By means of a magnetic field applied at right angles to the axis of the switch, the discharge can be confined to a small area. This considerably reduces the time required for the conditioning as well as reduces the necessary amount of vacuum melted material. A magnetic field of great intensity will disturb the focussing action and thus result in a faster deionization of the arc.

The use of parallel capacitors and resistors,[1,2] of course, also will improve the performance of a vacuum switch in reducing the recovery voltage.

Conclusion

It is believed that on the basis of the suggestions made here, a satisfactory working model capable of interrupting medium a-c or d-c loads can be constructed for test purposes.

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Semiconducting Shielding for A-C Power Cable

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ELECTRIC SHIELDING of a-c power cable is provided primarily for the purpose of directing and confining the electric field within the insulation so as to prevent the occurrence of unduly high electric stresses in dielectrically weak directions or regions of a cable. Martin Hochstadter laid down the principles governing such shielding in 1913, with particular application to impregnated paper insulated cables for high voltages.[1]

Such cable was not used widely during the following years, although the subject of shielding had been considered by many cable engineers in the United States. An excellent summary of shielding history and practice was presented in 1938 by L. F. Hickernell to the transmission and distribution committee of the Edison Electric Institute under the auspices of the Insulated Power Cable Engineers Association.[2]

During this time, various types of shielding, including the semiconducting type, were being considered. Semiconducting shielding may be defined as the practice of "partially" confining the electric field within the insulation by means of a semiconducting covering, its most important function being prevention of corona discharge. In 1928 cable with this form of shielding was described in an article by M. J. Lowenberg.[3] This article told of the failure of cotton-braided varnished-cambric cable, because of "arcs formed between the braid and the duct in which the cable was installed." Lowenberg then proceeded to tell how this trouble was eliminated by the substitution of a conducting cotton or asbestos braid for that of ordinary saturated cotton. The shielding effect of the braid, because of its partial- or semiconductivity, made possible the success of this venture.

Semiconducting shielding has been employed in a number of a-c cable installations in recent years, but an exact theory of its use has not been published. Consequently, most characteristics of this type of shielded cable have remained obscure, and the field to which semiconducting shielding may be applied safely has been limited unnecessarily.

It is hoped that the subject matter covered by this paper will assist in co-ordinating knowledge on the shielding of power cable, so that more effective use may be made of this electrical property.

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Theory of Semiconducting Shielding

While Lowenberg's paper dealt principally with the application of semiconducting shielding to varnished-cambric insulated station cables, probably its most important application is to rubber-insulated cables, as rubber is subjected to corona cutting from spark-generated ozone. The protective requirements of this type of installation may be met by semiconducting shielding.

Oscillographic observations of cable charging currents at ground points have verified the shielding effect of semiconducting cable coverings. Illustrative of this type of study are the oscillograms of Figure 1. They are typical of traces produced on the screen of a cathode-ray tube by the charging currents of three types of cable identical but for the electrical properties of their external coverings. Thus, the current traces of the cables having semiconducting and highly conducting shielding are smooth, and free of the high frequency components which are indicative of corona discharge. In contrast, unshielded cable under equal voltage conditions gives rise to a profusion of high frequency discharges. These oscillograms were obtained by carrying the 60-cycle charging current of the cable samples to ground through a 0.25 millihenry inductance, with the vertical plates of an oscillograph connected across the inductance.

To aid in interpreting the meaning of these observations, a theory of the electrical function performed by semiconducting cable coverings has been developed. This theory is based on an adaptation of the exact hyperbolic formulas for transmission lines to the circuit condi-

Figure 1. Charging current of shielded cable versus unshielded cable

tions present in a cable dielectric and semiconducting coverings. The simplified circuit of Figure 2 has been adopted as the electrical equivalent of cable insulation and outer coverings. The uniformly distributed capacity represents the dielectric, the conductance of which has been disregarded, its contribution to the admittance being negligible. The uniformly distributed resistance represents the semiconducting outer covering, the reactance of which also has been ignored, inasmuch as it has a comparatively minute effect on the impedance. The parallel capacity to ground likewise is neglected, as its operation reduces the current flow through the semiconducting covering by shunting a portion of the current, and, for practical purposes, this effect constitutes a factor of safety. Finally, the conductor is considered equivalent to the infinite ground in transmission calculations, inasmuch as for the cable lengths herein considered there is insignificant voltage drop along the conductor. The following formulas, expressed in co-ordinates adapted to measurement on the finished cable, were de-

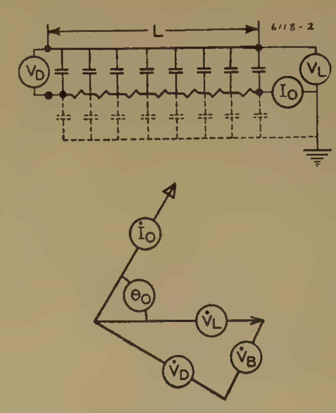

Figure 2. Equivalent circuit of cable insulation and outer semiconducting covering for single conductor cable

rived through these simplifying assumptions:[4]

$$V_B = V_L - \frac{V_L}{\cosh 0.001\sqrt{j(2\pi fcRL^2)}} \qquad (1)$$

$$V_D = \frac{V_L}{\cosh 0.001\sqrt{j(2\pi fcRL^2)}} \qquad (2)$$

$$I_0 = (10^{-9})(V_L)(L)\sqrt{j\left(\frac{2\pi fc}{RL^2}\right)} \times \frac{\sinh 0.001\sqrt{j(2\pi fcRL^2)}}{\cosh 0.001\sqrt{j(2\pi fcRL^2)}} \qquad (3)$$

$$\theta_0 = \left[\tan^{-1}\frac{\sin 0.1146\sqrt{\pi fcRL^2}}{\sinh 0.002\sqrt{\pi fcRL^2}}\right] + 45° \qquad (4)$$

where

V_L = vector line voltage to ground in volts
V_D = vector voltage drop across the insulation at a point midway between grounds
V_B = vector voltage drop lengthwise along the outer semiconducting covering, from the midpoint between two grounds to the ground points
I_0 = vector current in amperes at the ground point from the half-length of cable on one side of ground
θ_0 = power factor angle in degrees measured between the conductor and the ground point; this is the angle due to the power loss resulting from passing the charging current through the semiconducting shield to ground
f = frequency in cycles per second
c = capacity in micromicrofarads per inch length of cable
R = longitudinal resistance of the semiconducting covering in megohms per inch length of cable; this is the a-c resistance
L = one-half the distance between grounds in inches

Graphs for Use in Design of Semiconducting Shielded Cable

It will be noted that the foregoing formulas are functions of RL^2, RL^2 being a function of the a-c resistance of the semiconducting covering and the spacing between ground points. This fact has made feasible the calculation and drawing of graphs which have the common coordinate RL^2 and which encompass all probable cable constants. That this has been possible is very fortunate for, though simplified, the formulas still are comparatively unwieldy in applications to routine measurement and design; and it is quite difficult to appreciate fully the relation between the several electrical characteristics which they represent. The standard electric power frequency of 60 cycles per second has been employed in calculating the graphs, calculations being made for cables having capacities of 1, 5, 15, 30, and 50 micromicrofarads per inch length. Characteristics of cables possessing intermediate capacities may be obtained with sufficient accuracy for most practical applications by rough interpolation.

Figure 3 is a polar plot of the vector voltage relations at 60 cycles in a single conductor cable dielectric versus RL^2. Figure 4 is a semilogarithmic graph of the cable charging current versus the common abscissa, RL^2. Figure 5 likewise is a semilogarithmic graph with RL^2 as abscissa, and as ordinate the cotangent of the power factor angle, θ_0. Figure 6 is a similar graph of the maximum heat generation possible in watts in the inch of cable covering adjacent to the ground point, resulting from the passage of the charging current through the covering.

Figure 7 gives the thermal surface resistance in thermal ohms per inch length of cable for various cable diameters. These thermal resistance curves have been included because they are useful for determining the effect which semiconducting coverings may have on the thermal characteristics of a particular cable design, and because the data have been employed in checking the efficacy of the semiconducting shielding theory proposed. The curves have been derived from a paper published by S. J. Rosch.[5]

Examples in Use of Graphs

To illustrate the use of these graphs in cable design, consider an aerial cable supported on metal rings 20 inches apart, rated at 10,000 volts to ground, and characterized by a capacity of 15 micromicrofarads per inch and a semiconducting shield resistance to alternating current of two megohms per inch. For this case L equals 10 and RL^2 equals 200.

According to Figure 3, V_B, the maximum voltage drop longitudinally along the braid, would be equivalent to 55 per cent of the line voltage, that is, 5,500 volts. Moreover, V_D, the voltage impressed upon the insulation at the midpoint between grounds, would be reduced to 86 per cent of the line voltage.

Figure 4 predicts a charging current of 0.5 milliampere for the half length of cable on one side of a ground ring. As the semiconducting covering has a lengthwise resistance of two megohms per inch, a maximum voltage drop of one volt per mil would be developed by the passage of this charging current, a stress much too low to cause corona discharge.

Figure 5 shows that this cable would not behave as a pure capacity; instead it would have a cotangent θ_0 of approximately 0.37, which corresponds to an operating power factor of 35 per cent. Figure 6 indicates a loss of 0.5 watt in the inch of covering adjacent to the ground point.

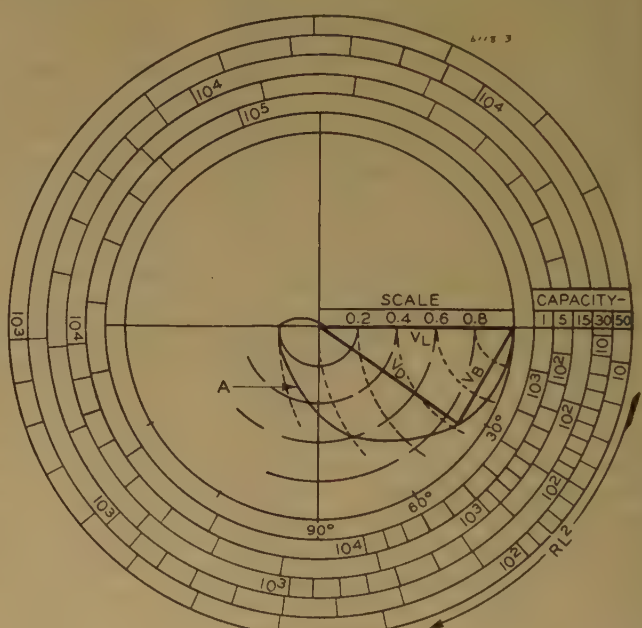

Figure 3. Dielectric vector voltages versus semiconducting shield resistance for single conductor cable

A—Locus of midpoint insulation and braid voltage vectors

If this cable has an over-all diameter of two inches, and is installed so that there is a free movement of air about it, the cable will possess a thermal surface resistance of approximately 24 thermal ohms per inch of cable length, according to curve A' of Figure 7. Since the temperature difference required to liberate heat from a cable surface is equal to the product of the watts times the surface resistance in ohms, a temperature rise of 12 degrees centigrade will be produced in the semiconducting covering adjacent to the

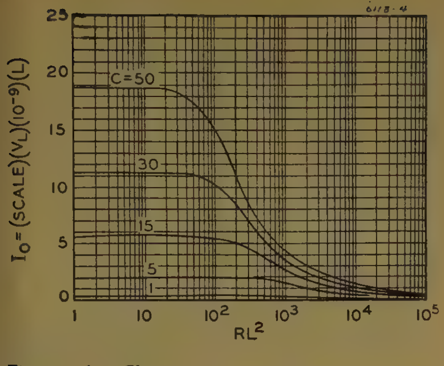

Figure 4. Charging current versus semiconducting shield resistance for single conductor cable

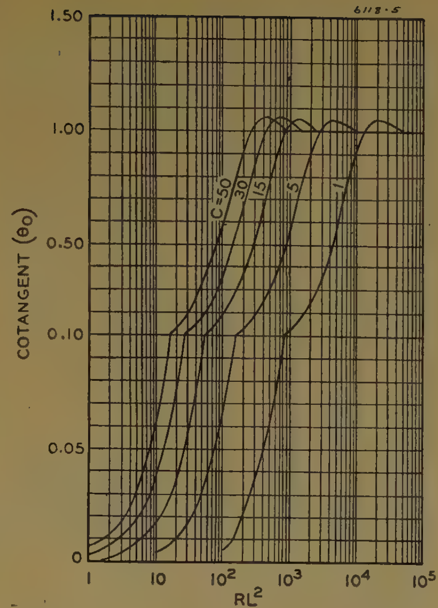

Figure 5. Power factor versus semiconducting shield resistance for single conductor cable

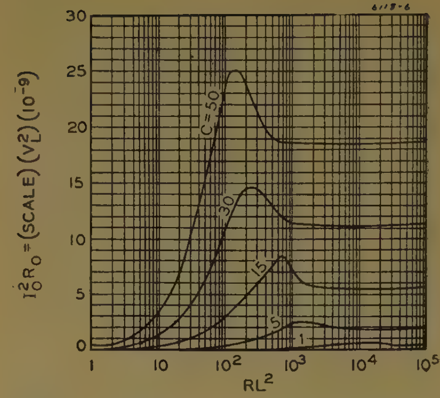

Figure 6. Watts loss versus semiconducting shield resistance for single conductor cable

ground point, because of the power loss resulting from passing the charging current through the semiconducting shield to ground.

Measuring Technique

In order to make available through the theory and graphs explained in the foregoing complete data relative to a particular sample of semiconducting shielded cable, seven constants must be measured. These are

1. The diameter over the outer coverings of the cable in inches, D.

2. The voltage applied for the measurements in volts, V_L.

3. The half-length in inches between ground loops, L.

4. The cotangent of the power factor angle of the combined semiconducting covering plus insulation, θ_2.

5. The ground point current corresponding to θ_2, I_0.

6. The cotangent of the insulation power factor angle, θ_1.

7. The capacity in micromicrofarads per inch length of cable, c.

In test procedure the first step is to measure the diameter, D, over the outer coverings. Next, the sample is prepared for the cotangent θ and ground point current measurements. The semiconducting covering is stripped from each end of the length, but is left intact on the center section. Wire loops are spaced on the active center section so as to form a number of π sections, similar to electrical models of long transmission lines or communication circuits. The number of sections and their length depend both upon the characteristics of the cable and the sensitivity of the bridge circuit employed for the measurements. Thus, in the analysis of a cable design having a low unit capacity and a high unit covering resistance, a larger number of sections of shorter length is necessary, than if the opposite be true. It is apparent that this aspect of the measurement requires an exercise of judgment and some experimenting before an efficient arrangement can be secured. In addition, the prepared sample must be placed on an insulated support, so that parallel capacity effects may be held to a minimum in accordance with the assumptions made in deriving the simplified formulas. With the sample so placed, all the wire loops except one are connected, through shielded leads, directly to the low tension side of a 60-cycle Schering bridge. The excepted loop is connected to the bridge through an a-c milliammeter or microammeter, so that the ground point current may be read. For the current measurement a ground loop adjacent to an end is preferable, because the current recorded at this point corresponds to the half-length current of the formulas and graphs. If the meter is connected to a loop in the center of the semiconducting section, twice this current is read, as it is collected from two half-lengths. The voltage now is applied and both the ground point current and cotangent of the power factor angle read. The half-length, L, corresponding to these measurements is recorded in inches, the voltage, V_L, in volts, the cotangent as that of θ_2, and the half-length current as I_0.

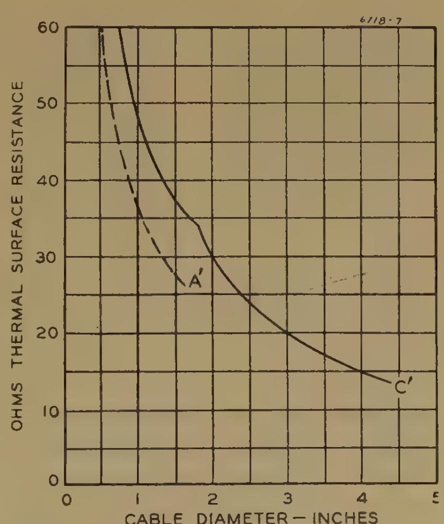

Figure 7. Surface resistance in thermal ohms per linear inch of cable for single conductor cable

A'—Braided cable in air
C'—National Electrical Manufacturers Association standard

Two measurements remain: the unit capacity, and the insulation power factor angle. To obtain these, the remaining semiconducting covering is stripped from the center section of the cable sample, and replaced with a highly conducting metal foil wrapping. Then the cotangent of the insulating power factor angle, θ_1, is measured, as is the capacity of the length in micromicrofarads. The capacityin micromicrofarads per inch, obtained from this measurement, is the unit capacity, c, employed in the formulas and graphs described in the foregoing.

The first step in employing these measured data to make an analysis of the characteristics of the semiconducting covering is to obtain the value of cotangent θ_0. This may be done by subtracting θ_1 from θ_2, the difference angle being θ_0, this procedure is in accordance with one of the original assumptions in deriving the theory, namely, that the insulation conductance was ignored so as to aid in simplifying the formulas and calculations. Having found the value of cotangent θ_0, and knowing the capacity of the cable in micromicrofarads per inch of length, the corresponding RL^2 may be obtained from Figure 5. In turn, the unit resistance of

Table I. Comparison of Calculated and Experimental Data

Sample	Cable Specification: D	V_L[a]	L	R[a]	C[a]	Measured: V_B	V_D	I_0[c]	T_0	Calculated: V_B	V_D	I_0	T_0
A	1.62	5,000	39	0.01	8	120[b]		0.6×10^{-3}		125		0.63×10^{-3}	
		72,000	39						21				19
B		5,000	22	0.26	5.6	115[b]		0.24×10^{-3}		750		0.24×10^{-3}	
C		111	10	2.5	11.3	68[d]	100[d]			60	99		
		10,000	5	3.6	11.3			0.22×10^{-3}				0.21×10^{-3}	
D		2,500	5	4.4	10			0.048×10^{-3}				0.045×10^{-3}	

[a] = value of constants obtained from calculations based on 60-cycle Schering bridge measurements.
[b] = measured with vacuum tube voltmeter
[c] = measured with rectifier type milliammeter
[d] = null-type measurement, Figure 8

the semiconducting covering in megohms per inch of cable length may be obtained by dividing the value of RL^2 by the square of the half-length used in the cotangent θ_2 measurements. Having established the value of R, it becomes comparatively simple to predict the performance of the cable design for any ground point spacing by calculating the particular RL^2 and applying this to Figures 3 through 7.

Experimental Confirmation of Theory

Table I contains data obtained from cable designs which were selected to test the correctness of the theory. The choice of samples represents a range of 60-cycle semiconducting covering resistance from approximately 0.01 megohm to slightly over four megohms per inch of cable length. The agreement between values calculated from the cotangent measurements and those measured directly is very good for this type of measurement and thereby certify the practical worth of the theory.

Inasmuch as some interesting problems in measuring technique presented themselves in the course of this experimental work, it has been thought worthwhile to explain them in detail for their possible application to similar measurements in the field of dielectric research. Many measurements were made which did not fit in as well as was expected with the calculated theory. This was true particularly of the readings taken of the voltage built up lengthwise along the semiconducting covering by the flow of the cable charging current. For example, the measured V_B of sample A agreed very well with that calculated, but that of sample B did not. Both these voltages were read with a vacuum tube voltmeter, which was supposed to approximate a zero energy device. This it does for most measurements to which it is applied, but in the case of sample B, which possessed a covering of medium high resistance, the parallel capacity to ground of the voltmeter pickup prod was sufficient to constitute a partial ground. As a result, a considerable portion of the cable charging current passed through this partial ground and therefore was not available to cause a voltage rise along the covering. This phenomenon is illustrative of the manner in which parallel capacity to ducts and conduits affects the characteristics of semiconducting shielded cable.

The vacuum tube voltmeter was still more ineffective in attempts to measure the voltage characteristics of sample C, which was characterized by a high-resistance "semiconducting" covering. It was necessary to employ a true zero energy measuring technique for these readings. The circuit of Figure 8 shows the method by which this was accomplished. The voltage built up on the cable covering under test was balanced at

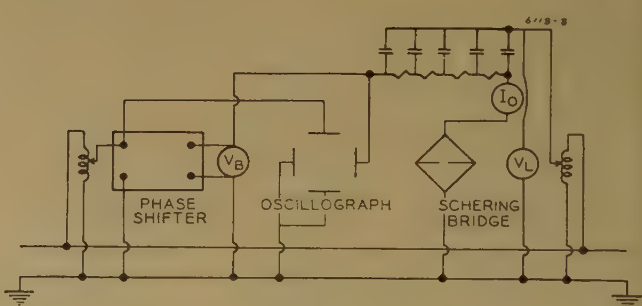

Figure 8. Null point method of measuring phase and value of voltage on the semiconducting covering of semiconducting shielded cable

a midpoint between grounds by a voltage applied through a phase shifter. The magnitude of this applied voltage was varied by means of an autotransformer placed in the primary winding of the phase shifter, and the value of the voltage output of the phase shifter was read by a voltmeter. The phase of the voltage applied to the covering as related to the line voltage impressed on the cable conductor was indicated by connecting a cathode-ray oscillograph as shown to form Lissajous figures.[6] A Schering bridge was used to indicate balance; thus, if the braid voltage were not balanced by the voltage from the second circuit, the latter would constitute a partial ground and cotangent θ would be lowered as a result of the length between grounds in effect being shortened. If the braid voltage were overbalanced, cotangent θ would be raised; but the same cotangent would mean a null balance, that is, V equal to V.

Discussion

Semiconducting shielding, like high conductivity shielding, requires adequate grounding. If contact resistance or impedance is too high, arcing will occur at ground points with attendant destructive thermal and chemical effects.

Power loss in a semiconducting shield also may be, under certain conditions, sufficiently great to reduce the current carrying capacity of a cable markedly. Thus, consider a 2,000,000 circular mil cable, with a unit capacity of 15 micro-microfarads per inch, an over-all diameter of 2.7 inches, operated at 10,000 volts to ground. If RL^2 equaled 600, the power loss in the inch of semiconducting shield adjacent to the ground point would be 0.8 watt. Inasmuch as the thermal surface resistance of this cable is 23 thermal ohms per inch, a hot spot would result with a temperature rise equal to the product of watts and thermal ohms, or 18 degrees centigrade above normal. This temperature rise would result entirely from the passage of charging current, and would have to be superimposed upon that resulting from the load current.

Conclusions

A theory of semiconducting shielding has been presented which may be used as a guide in the application of this type of shield to electric power cable.

The electrical resistance required of a semiconducting shield for a particular cable installation is determined by several interrelated factors: the maximum spacing between ground points, the cable capacity per unit length, the line voltage, and the over-all diameter of the cable.

To be noted particularly is the fact that all pertinent electrical characteristics are functions of the product of the shield resistance and the square of half the ground point spacing.

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Automatic Control of Large Synchronous Condensers

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Synopsis: The considerations involved in the design of an automatic synchronous condenser control are discussed in this paper with an aim toward helping the designer and user of synchronous condenser equipment obtain modern and complete facilities in his design. A typical push-button-operated control for a large synchronous condenser is described and the diagrams involved are shown. The description includes a detailed résumé of starting sequence, protective device operation, and automatic control of voltage and load. Emphasis is put on the adaptability of push-button control to full automatic operation or remote operation by direct or supervisory control.

WHILE there is no clear line of demarcation between large and small synchronous condensers, it will best serve the purposes of this discussion to define large synchronous condensers as those having a rating of 5,000 kva or above. Numerically, there are about as many synchronous condensers below the 5 000-kva rating as above it, but the connected capacity of the large machines represents nine-tenths of the total capacity and the small machines only one-tenth. Hydrogen cooling appears on machines above 15,000-kva rating and at present will be found on about 12 per cent of these machines. There is reason to believe that hydrogen cooling will become more general in this class.

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Automatic control for condensers in the larger sizes offers advantages that are attractive to users of these machines. The starting procedure, which begins with the acceleration of the rotor and involves field application at the time that synchronous speed is reached and ends with applying full voltage to the machine stator, is a routine sequence that involves several operations that must be properly timed. The timing of these operations as well as the checking of various mechanical and electrical conditions during the starting sequence is a duty that can be more efficiently handled automatically than manually. Besides this very practical consideration of accurate procedure, there are other reasons for adapting automatic control. The time consumed in going through the starting procedure will be a minimum, and the operating personnel can be reduced in number or skill. These two features are of interest to operating supervisors who bear the responsibility for efficient and punctual operations in the system. Automatic operation, which is manually initiated and terminated, will have as its primary element a manually operated push-button or control switch. It is not difficult to extend this control to full automatic operation wherein the machine is started by undervoltage conditions and removed when machine load drops. Remote control of starting and stopping is another alternative. Such control simply is an extension of the local manual control over a direct wire or as a point on a supervisory control system.

Apparatus Associated With Large Synchronous Condensers

There are two general methods of obtaining starting torque. One method employs an induction motor directly connected to the condenser shaft as a source of accelerating torque. The second is a self-starting method wherein the reduced voltage is applied to the stator and the rotor accelerates on torque from the damping bars. Which of these two methods is to be used is determined by system conditions.

Should the system demand minimum disturbance, the induction motor should be used. If, however, it can tolerate the power swings involved in low voltage starting, this method will be found to be more economical in both equipment and space requirements.

Low voltage starting may be effected by using an autotransformer, tapped power transformer, series reactor, or a series-parallel stator connection scheme. The autotransformer scheme is the most generally used and is the basis of the control discussed in the following pages.

Two autotransformer connection schemes are in general use. Figure 1 shows the two schemes. They differ in the location of the starting breaker. The choice between the two is based upon breaker costs and economy of installation. Satisfactory performance is obtained with either connection, and neither has any operating advantage over the other.

Excitation for the field can be obtained from either a direct-connected exciter or a separately driven exciter. The latter is employed with hydrogen-cooled machines when it is desired to keep the d-c machine and its commutator out of the hydrogen chamber. The main exciter will be equipped with a main and differential field, the latter being used for overcoming the exciter residual so that the low voltages required for lagging capacity may be obtained. The main and differential field will be under the control of the voltage regulator system whose motor-driven rheostat and field forcing resistors will alter excitation to meet changing demands for reactive power output of the condenser. A pilot exciter on the same shaft as the main exciter will furnish excitation current for the main exciter fields. A field cubicle containing the field breaker and discharge resistor completes the excitation equipment.

The larger condensers will be equipped with bearing lift pumps that must be used as a means of floating the shaft on a film of oil prior to starting. There will

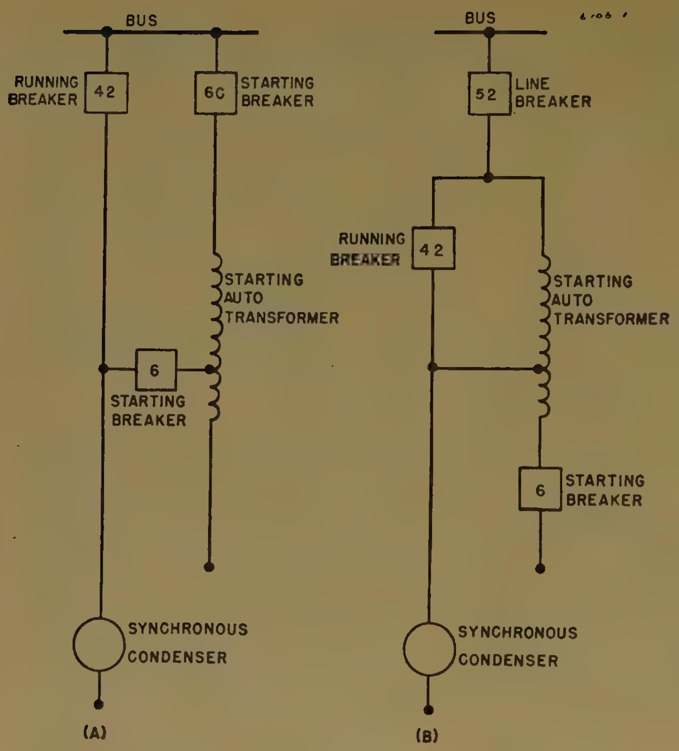


Figure 1. Synchronous condenser connections to autotransformer

be oil pumps employed in some cases for circulating lubricating oil in the bearings that must operate continuously as long as the motor is running. On other machines, oil rings provide lubricating oil circulation without the need for auxiliary oil pumps.

The instruments, relays, control switches, and associated devices will require switchboard mounting. Figures 2 and 3 show one section of a tunnel-type switchboard designed for the control of a 25,000-kva synchronous condenser. This type of construction has been adopted widely because the control switches and instruments are very accessible and constantly observable on the front panel of the switchboard, while the relays are easily attended on the back panel of the board. All wiring, terminal blocks, and fuses are concealed and protected within the tunnel housing. This area of the switchboard can be reached by entering a door at either end of the switchboard assembly. Drawout relays with built in test switches offer flexibility of handling when maintaining the relays.

Starting Characteristics of Synchronous Condensers

When a condenser is brought up to speed with a direct-connected induction motor, the problems encountered in putting it on the line are exactly the same as those experienced when a generator is put on a line. The machine must be brought to synchronous speed, its voltage matched to the line voltage and synchronized with it. At the time when these conditions exist, the running breaker may be closed. The disturbance caused by closing the running breaker will be negligible if voltage and phase are matched closely.

When self-starting on reduced voltage is employed, the least disturbance is encountered if the field is applied before full voltage is put on the stator. Figure 5 shows the "V" curves for a typical condenser. The lower curve indicates the line current to be expected for various values of field excitation when operating the machine at 22.5 per cent rated voltage. The upper curve indicates line current variation with field excitation when the machine is operating at full line voltage. The intersection of these two curves gives a value of field current which may be used during starting if the leading reactive load during starting is to be the same before and after transfer from starting to running voltage.

The sequence of events in a typical start is shown by the graph in Figure 6. This shows the variation of reactive

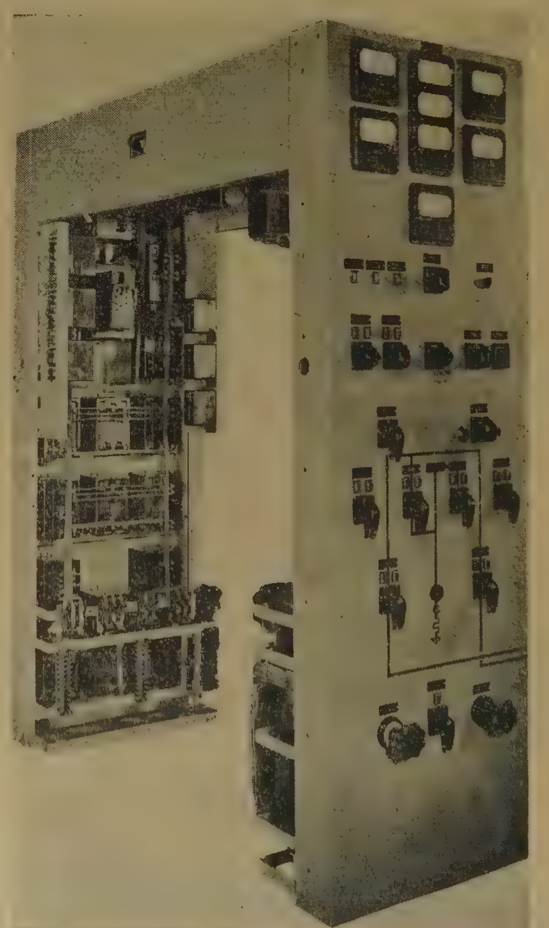


Figure 2. Tunnel-type switchboard section designed for synchronous condenser control

Figure 3 (right). Rear panel relay arrangement of switchboard shown in Figure 2

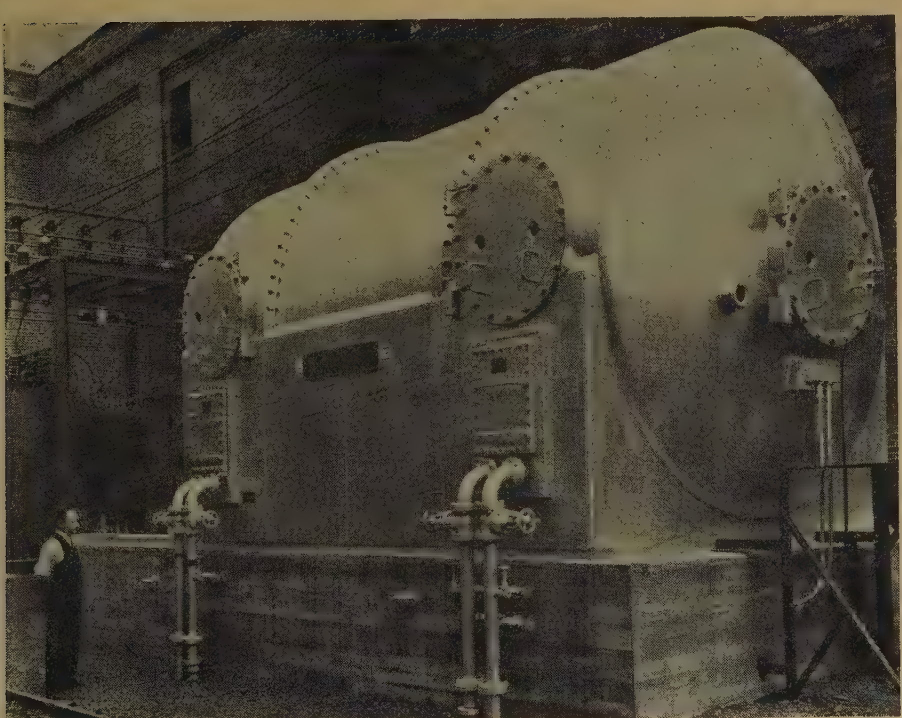
power, as seen by a graphic varmeter when the machine is accelerated and put on the line. The initial reactive inrush is lagging and diminishes in value as the machine accelerates. At half speed a slight drop is noted, but the decline continues smoothly until synchronous speed is approached. Hunting appears as a wavy trace in the graph and as synchronous speed is reached, the reactive power drops to a low steady value. When the field is applied, the reactive power becomes leading, building up in magnitude as the field builds up. The field eventually stabilizes and the machine is operated for a brief period under these conditions to assure that hunting has subsided. The transfer from starting to running voltage is made at this point. Although the actual transfer requires only a fraction of a second its effect is to reduce field flux to a low value. A subsequent delay is encountered in rebuilding the field. During the transfer, the field current will be dropped to nearly zero as the induced voltage in the field winding opposes the exciter voltage. The reactive power changes sharply during transfer from leading to lagging and returns slowly to leading again as field flux reappears.

The disturbance on a system caused by a synchronous condenser coming on the line is the net effect of the various reactive power swings encountered in the period extending from the time of initial inrush to final stabilized operation at full line voltage. The best operating point can be determined only by evaluating the effect of the individual disturbances during this period on the system to which the condenser is applied. The field excitation determined by the intersection of the "V" curves in Figure 5 will give satisfactory operation in many cases. A decrease in field current will increase the severity of the lagging reactive power peak encountered at the time of transfer. An increase in excitation will decrease this peak, but it also will bring the condenser on the line at a higher load, at both the starting and running voltage.

Table I. Device Table for Schematic Diagrams

1 = start-stop push button
4 = master relay (multicontact type)
6 = starting breaker
14 = locked rotor protective relay (slow release telephone type type)
14X1 = pendulum relay (telephone type)
14Y1 = counting chain relay (telephone type)
14Y2 = counting chain relay (telephone type)
14Y3 = counting chain relay (telephone type)
14Z = pulsing relay (telephone type)
18 = acceleration relay (current operated induction disk type)
18X = auxiliary relay (solenoid contactor)
26 = thermal relay
26H = hydrogen temperature in machine
30 = annunciator relay
30-1, 30-2, and so forth = individual drops
30X = auxiliary relay (electrically operated, manually reset)
30Y = bell alarm cutoff relay (solenoid contactor type)
38 = bearing thermal relay
40 = field failure relay (solenoid contactor type)
41 = field breaker
42 = running breaker
46 = current balance relay (induction disk type)
47 = undervoltage, single and reversed phase relay (induction disk type)
48 = incomplete sequence relay (a-c timing relay)
49 = thermal overload relay (thermal lag—thermostatic relay)
51 = inverse time overcurrent relay (induction disk type)
52 = line breaker
53 = exciter voltage relay (instantaneous solenoid type)
56 = field application relay (a-c timing relay)
T_1 = short time contact
T_2 = long time contact
59 = a-c overvoltage relay (induction disk type)
63 = pressure relay
63H = hydrogen pressure in machine
63HB = hydrogen pressure in bottles
63QF = oil pressure in front bearing
63QR = oil pressure in rear bearing
63QX = auxiliary oil pressure relay (solenoid contactor type)
63W = cooling-water pressure
63WM = water detector relay
86 = lockout relay (electrically operated, manually reset)
87 = differential relay (induction disk type)
88 = auxiliary motor
88E = exciter motor
88EX = exciter motor breaker
88W = cooling-water valve motor
88WX = cooling-water valve motor relay
88Q = oil lift pump motor
88QX = oil lift pump motor relay
90 = voltage regulator
90C = current regulator operating coil
90CX = a-c timing relay (induction disk type)
90CY = a-c fault detector relay (instantaneous solenoid type)
90CHR = antihunt coil—current element—raise
90CHL = antihunt coil—current element—lower
90CZ = auxiliary relay (multicontact type)
AR = quick-raise contact (voltage element)
AR' = quick-raise contact (current element)
AL = quick-lower contact (voltage element)
CR = raise contact (current element)
CL = lower contact (current element)
CSR = control switch raise
CSL = control switch lower
I = contact closed in indicating position of regulator transfer switch
LC = lowering contactor
LS = limit switch
M = contact closed in manual position of regulator transfer switch
NR = normal-raise contactor
NL = normal-lower contactor
NH = antihunt coil—normal raise and lower contacts
QR = quick-raise contactor
QL = quick-lower contactor
QH = antihunt coil—quick-raise and -lower contacts
R = contact closed in regulating position of regulator transfer switch
RL = red light indicating raise required
LL = green light indicating lower required
VR = raise contact (voltage element)
VL = lower contact (voltage element)

Figure 4. A 15,000-kva hydrogen-cooled outdoor synchronous condenser



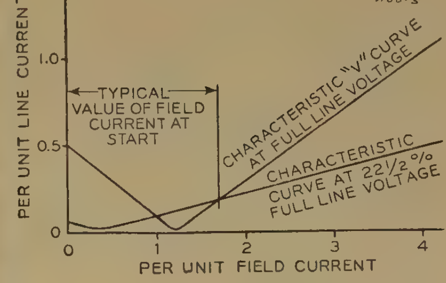


Figure 5. Characteristic curves for a typical synchronous condenser

An alternative method of starting suggests itself; that of accelerating the machine on low voltage, applying full voltage, and then applying field at the no-load value. The inrush of reactive power is far greater in this case when full voltage is applied than in the case where field is on before full voltage is applied. With this method, the phase and magnitude of the internal voltage of the condenser are less favorably matched to line voltage phase and magnitude when transfer is made than would be the case if the field were excited. Consequently, the inrush is greater with this method than with the method where field is applied before transfer is made.

Automatic Sequence for Controlling Synchronous Condenser

The following description applies to the control of a self-starting hydrogen-cooled machine operating from an autotransformer. Figure 7 shows the single line diagram of this control and Figure 8, the control diagram. This is a typical scheme and would be altered in practice by changes in details to suit each particular application. The functional sequences are as follows:

Push-Button Starting

1. Operate push button.

2. Start incomplete sequence timing relay. Failure to complete the starting sequence in a definite length of time will cause lockout.

3. Check position of control elements.
(*a*). Check to assure that line breaker is open.
(*b*). Check to assure that starting breaker is open.
(*c*). Check to assure that running breaker is open.
(*d*). Check to assure that field breaker is open.
(*e*). Check to assure that motor-operated rheostat is in pre-set position.
(*f*). Check to assure that lockout relay has been reset.
(*g*). Check to assure that proper voltage and phase rotation exists on the a-c line.
(*h*). Check to assure that station battery voltage is at an operative value.

4. Start auxiliaries.
(*a*). Start oil lift pumps.
(*b*). Start separate exciter.
(*c*). Start flow of bearing cooling water.

5. Check operation of auxiliaries.
(*a*). Check oil lift pump pressure.
(*b*). Check exciter voltage.
(*c*). Check cooling water pressure (or flow) to hydrogen cooler and bearings.

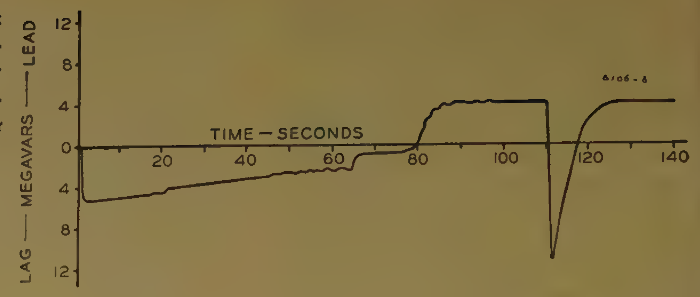
6. Close starting breaker.

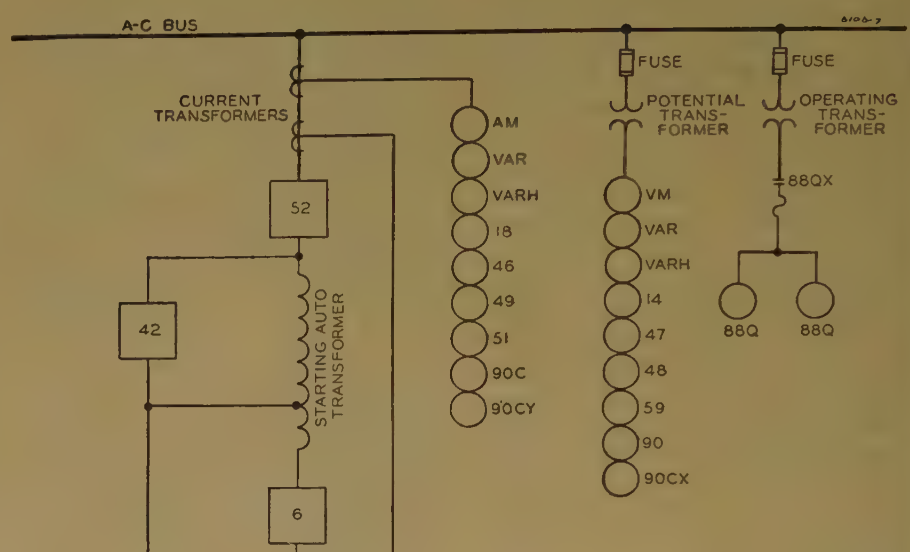
7. Check closing of starting breaker.

8. Close line breaker.

9. Check for machine rotation. Failure to rotate will cause lockout.

Figure 6. Graphic varmeter trace of reactive power drain by a 25,000-kva synchronous condenser during starting





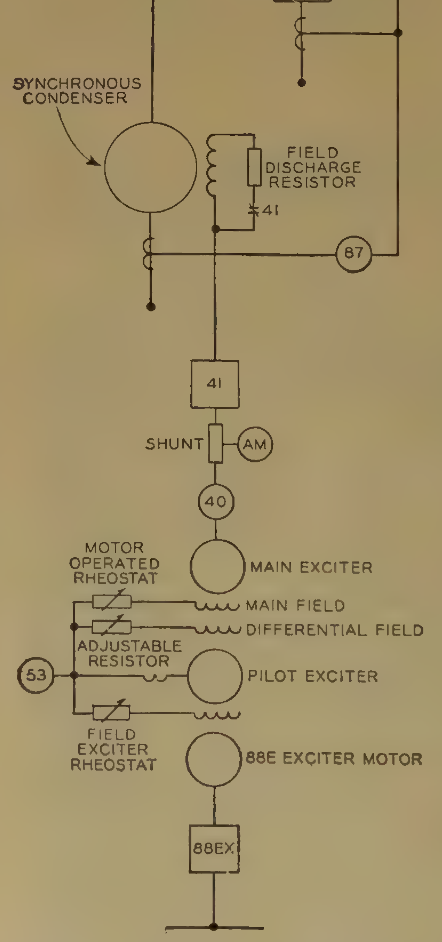


Figure 7. Single-line diagram of a typical synchronous condenser and associated equipment

10. Start field application timing relay when machine reaches approximately two-thirds of full speed. This relay ultimately applies field and later trips starting breaker and closes running breaker.

11. Apply field by closing field breaker after field application relay closes short time contact.

12. Trip starting breaker after field application relay closes long time contact.

13. Check opening of starting breaker.

14. Close running breaker.

15. Connect voltage regulator.

16. Shut down oil lift pumps.

17. Connect contacts of field failure relay

Push Button Stopping

Note: The machine's load should be reduced to zero before taking it off the line. This must be done manually unless the control is fully automatic. In this case the loss of load precedes the stopping sequence.

1. Open line breaker.

2. Open running breaker.

3. Check to assure that running breaker is open

4. Open field breaker.

5. Check to assure that field breaker is open.

6. Open exciter motor breaker.

7. Drive motor-operated rheostat to starting position.

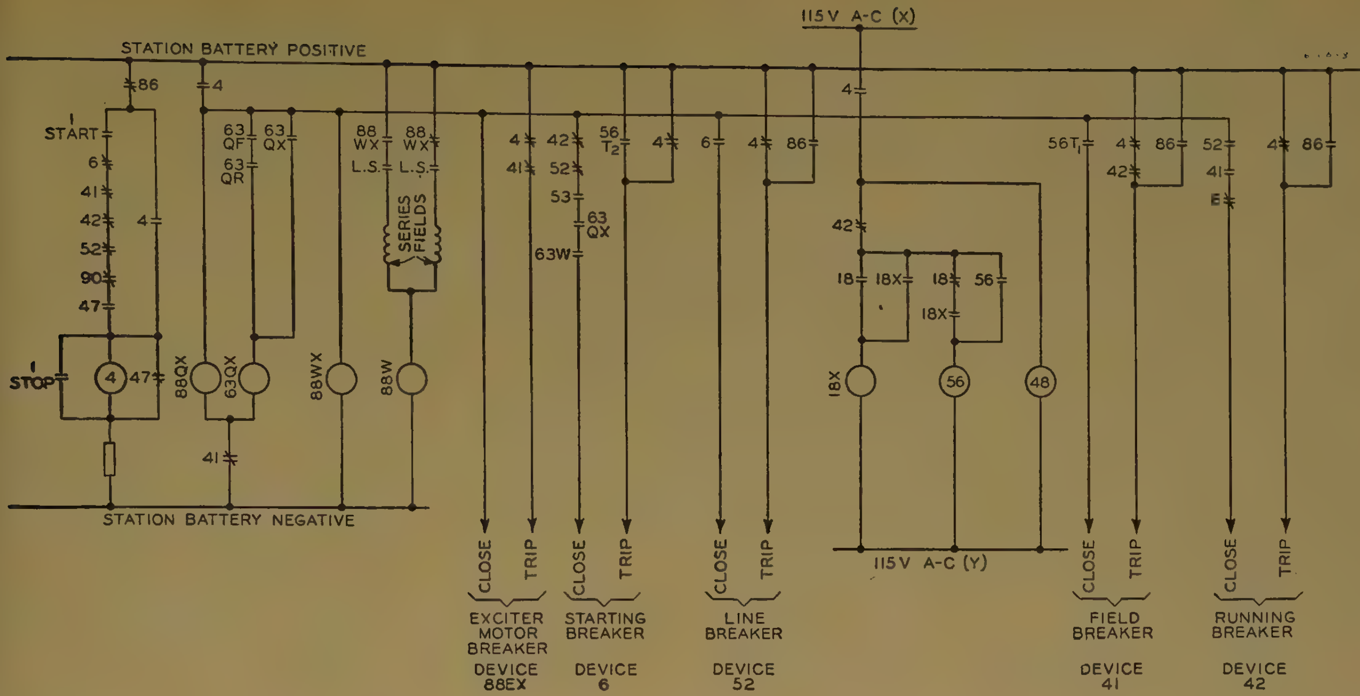


Figure 8. Control-relay diagram for automatic control of a synchronous condenser

8. Shut down accessories: Hydrogen and bearing cooling water.

EMERGENCY STOPPING

1. Open line breaker.
2. Open running breaker.
3. Open field breaker.
4. Check to assure that field breaker is open.
5. Open exciter motor breaker.
6. Drive motor operated rheostat to starting position.
7. Shutdown accessories: hydrogen and bearing cooling water.

Voltage and Current Regulator

After the machine is on the line, voltage regulation of the bus to which it is connected can be delegated to a voltage regulator. This regulator increases or decreases field current as the bus voltage tends to decrease or increase. The machine, however, is limited by its capacity in the amount that it can correct line voltage. If it cannot raise the line voltage, it will overload itself in the attempt unless something is done to limit the reactive load that it takes. For this reason, a current regulator is added to the voltage regulator scheme on condensers. If the voltage regulator tends to raise excitation to a point where the line current would exceed the maximum permissible value, then the current regulator makes the voltage regulator ineffective. If the line current exceeds this value for any reason, then the current regulator lowers the excitation to a safe value.

During fault conditions there would be a tendency for the current regulator to take immediate control and reduce excitation. Since this is a transient condition, a timing relay scheme is employed that holds the current regulator ineffective for a short time during the fault. If the fault persists, the current regulator begins to operate and reduces excitation.

The bus voltage value that the condenser maintains can be selected by adjustment of the voltage regulating rheostat in the voltage regulator circuit.

In some applications it is the purpose of the synchronous condenser to keep the power factor of a load constant even though this load may change rapidly and its power factor vary widely. Electric arc furnaces draw this type of load. In these cases it is necessary to employ a power factor regulator instead of a voltage regulator. The power factor regulator operates on load current and voltage and is sensitive to changes in the phase angle between the two. If the power factor drops below the desired value, a contact closes, causing increased condenser excitation. A raise in power factor above the regulating value decreases excitation. A current-limiting element must be used with the power factor regulator. However, quick raise and quick lower devices are not necessary, nor are the devices employed with the current regulator for rendering it ineffective during faults.

Protecting Synchronous Condenser

The abnormal conditions which may develop in the synchronous condenser, its control, and auxiliaries may be segregated into two classes. The first group embraces those troubles which are dangerous because of their damaging effect on the machine, auxiliaries, or connected lines. These demand immediate protective action and in general call for a complete shutdown and lockout. The second group includes the abnormal conditions that cannot be immediately responsible for damage or disturbance but demand attention before more serious consequential conditions arise. These faults demand immediate annunciation but do not require shutdown.

In the first group are the following conditions, conveniently termed lockout faults:

1. Locked rotor. Machine fails to rotate when starting voltage is applied.
2. Bearing overtemperature.
   (*a*). Condenser.
   (*b*). Exciter.
3. Field failure.
4. Overcurrent.
   (*a*). Instantaneous.
   (*b*). Inverse time delayed.
   (*c*). Long time overload.
5. Overvoltage.
6. Internal fault in machine (differential relay operation).
7. Phase current unbalance.
8. Failure to complete starting sequence.

The second group, which may be

termed annunciating faults, are as follows:

1. Overtemperature of hydrogen.
(a). In main chamber of machine.
(b). In collector chamber of machine.

2. Underpressure in hydrogen bottles.

3. Underpressure in machine.

4. Presence of water in water collectors beneath machine.

Full Automatic Control

The start contact in Figure 8 can be replaced by a delayed make-contact of an undervoltage relay. The delay eliminates starting on passing conditions of undervoltage and assures that the condenser will be started only when voltage is low and remains low for a definite period of time.

The stop contact can be replaced by an underload contact which closes in a definite period of time after the condenser load drops to a low value.

The voltage regulator in the full automatic station controls the load on the condenser. Since the voltage-adjusting rheostat will be pre-set, the regulator will alter excitation as required to maintain the bus voltage corresponding to this setting. The bus voltage which it will tend to maintain can be altered by changing the voltage-adjusting rheostat manually.

Supervisory Control

The control of starting and stopping is only one of the functions which must be considered when applying supervisory control to the condenser station. There also must be considered the supervision of protective devices, telemetering of the electrical quantities involved, and remote operation of the voltage adjusting rheostat.

A practical operating scheme would provide continuous telemetering of kilovars. The supervisory control would assign two points to the condenser control. One point would permit starting or stopping of the condenser. Following a start operation an indicating light would show when the master relay picked up and another light would show when the running breaker closed. A third light indication would follow the stop operation. The second supervisory point would operate the voltage-ajusting rheostat. Changes in kilovar load resulting from operation of the voltage-adjusting rheostat would be observed on the kilovar telemeter by the dispatcher. The supervisory control may be operated over a metallic pair or a carrier current channel.

Detailed Description of Typical Scheme

The synchronous condenser connections of Figure 7 can be considered typical for a hydrogen-cooled machine of large size. Instrumentation is simple. Line voltage, line current, kilovars, and kilovar-hours are the a-c quantities of interest. Field current, main exciter voltage, and pilot exciter voltage are the d-c quantities of interest. A temperature indicator connected to the exploring coils in the machine, although not shown schematically here, is also desirable on the switchboard.

The main exciter motor operates from the station service 440 volts. An operating transformer supplies 230 volts a-c to the oil lift pump motors. A potential transformer on the high voltage bus supplies the instruments and relays. A double secondary current transformer on the line side of the condenser breaker is connected so that one secondary operates instruments and relays and the other secondary operates the differential relay. During starting, the current in the condenser windings will be several times the line current, and a third set of current transformers may be employed in the autotransformer neutral to balance the currents on the two sides of the differential relay during a normal start. An alternative method is to use a differential relay that is spring-restrained from operation at the values of current involved during starting. This adjustment is normally within the range of conventional differential relays.

In Figure 8 the progression of the starting sequence can be followed by moving from left to right on the diagram. If, as will be seen from the circuit to coil 4, the lockout relay is reset, the starting push-button depressed, all breakers out, and the motor-operated exciter field rheostat in the proper position for starting, normal voltage conditions existing, relay 4 will pick up and seal in around all contacts except the lockout relay contacts. It will remain picked up unless short-circuited by the stop push-button contact or the reverse phase and undervoltage relay, or de-energized by the opening of the lockout relay.

Make contacts of relay 4 set up circuits to all the devices operated in the starting sequence. Linestarter 88*QX* is operated immediately after relay 4 picks up to start operation of the oil lift pumps. Likewise, the cooling water valve motor relay 88*WX* immediately is picked up to cause opening of the cooling water valve. At the same time, the exciter motor breaker is closed to start the exciter and the incomplete sequence relay timer is started.

When lift pump oil pressure builds up, relay 63*QX* is picked up, seals itself in, and is released later when the field breaker closes. The contacts of relay 63*QX* in series with a cooling water pressure contact and an exciter voltage relay contact initiate the closing of the starting breaker, device 6. A contact of 42 is in this circuit to assure that under no circumstances can breaker 6 be closed when breaker 42 is closed.

The closing of starting breaker, device 6, initiates the closing of the line breaker, device 52. After the line breaker closes, the condenser is on the line at low voltage, and it will draw current that will tend to cause relay 18 to close its front contacts. Relay 18 is a current-operated induction disk relay, and as the machine accelerates, it drops out on the reduced current, eventually closing its back contact. When 18 closes its front contact, it picks up relay 18*X*, which seals itself in. A contact of 18*X* in series with the back contact of 18 picks up and seals in timer 56.

At the end of a definite timing period, relay 56 closes the field breaker, device 41. Further operation of timer 56 causes starting breaker, device 6, to trip. The condition that the line and field breakers are closed and breaker 6 is open sets up a circuit to close breaker 42, thereby putting the machine on the line at full voltage.

The voltage regulator shown in Figure 9 is brought into service by the closing of breaker 42, provided the control switch whose contacts are shown as *RI* is in either the regulating or indicating position. This switch has "manual," "indicating," and "regulating" positions. In the "manual" or "indicating" position, the motor-operated rheostat may be raised or lowered by hand through the contacts *CSR* and *CSL* of a control switch. In the "indicating" position, the voltage regulator performs all of its functions except that it does not move the motor-operated rheostat. In the "regulating" position, the voltage regulator is in full operation.

The d-c motor which drives the motor-operated rheostat has a split series field. Energization of the armature in series with one field will raise excitation; the other will lower. Limit switches prevent excessive travel of the mechanism. Cam switches are connected in the rheostat motor circuit to drive the rheostat to the starting position after the running

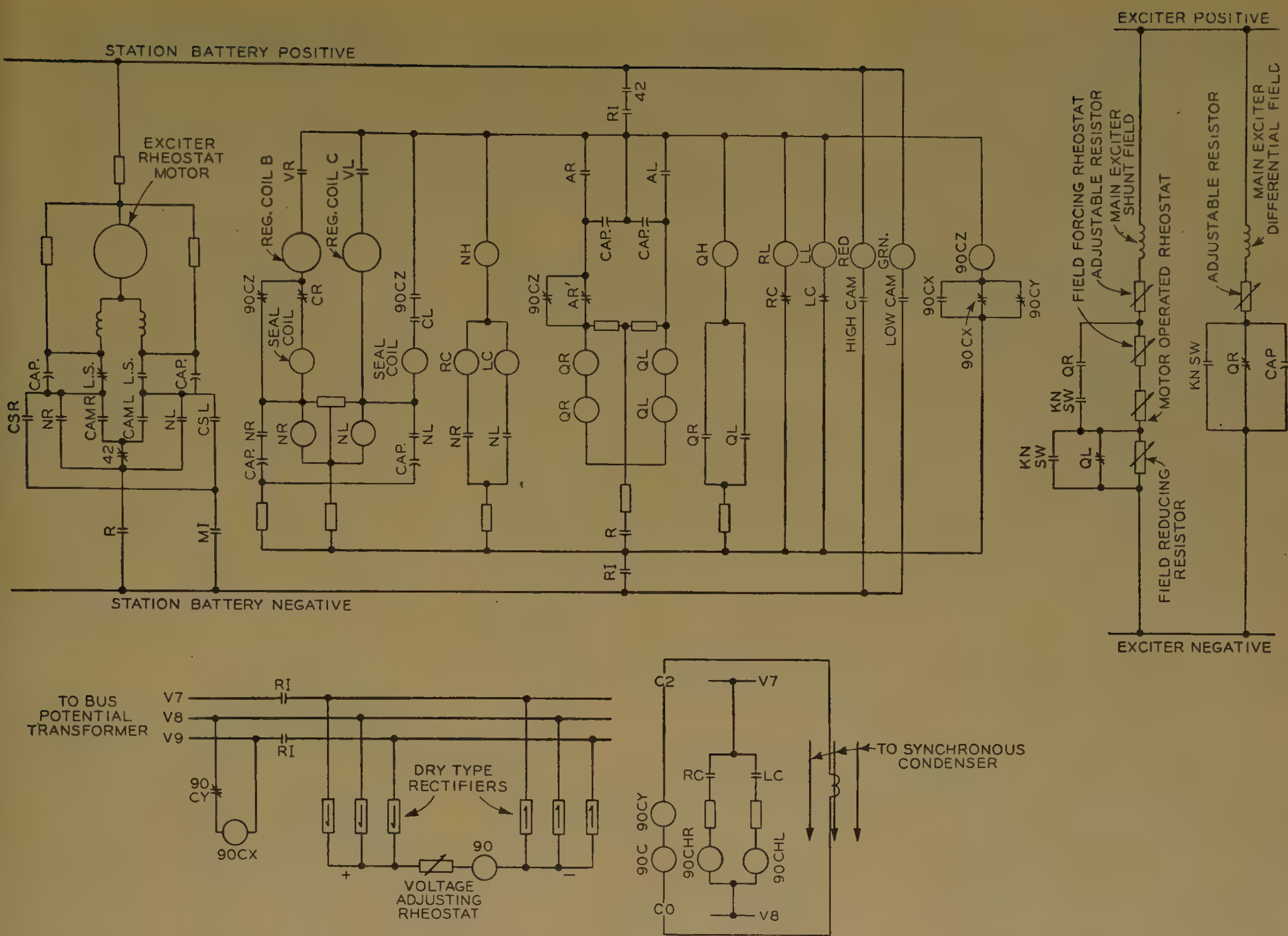


Figure 9. Voltage and current regulator connections used with synchronous condenser control

breaker closes on a shutdown. Contactors *NR* and *NL* control the rheostat under direction of the voltage regulator.

An indication comes from the voltage regulator element that a raise in excitation is necessary by the closing of contact *VR* in series with regulator coil *B*. Contactor *NR* is picked up by this action, and it not only operates the motor-operated rheostat, but picks up relay *RC* and antihunt coil *NH*. The latter functions mechanically to open contact *VR* even before any rise in voltage is felt on the bus. *VR* is permitted to close if, after a time delay, an adequate rise in bus voltage is not experienced. Relay *RC* provides back contacts to operate the indicating light *RL* and front contacts to operate the current element antihunt coil 90*CHR*. Similarly, a demand for reduction in voltage involves contact *VL*, regulator coil *C*, contactor *NL*, relay *LC*, and antihunt coil *NH*.

Sudden wide changes in voltage demand more rapid changes in excitation than normal operation through the motor-operated rheostat will provide. In this case, contacts *AR* and *AL* operate the quick-raise contactor *QR* or quick-lower contactor *QL*, as the case may be. These contactors energize the antihunt coil *QH* and also switch the field forcing or field reducing resistors into the circuit. In the event of a demand for quick raise, *QR* de-energizes the differential field also.

The current regulating coil is effective during moderate overcurrent conditions and also during fault conditions of extended duration. Normally the time delay relay 90*CX* is energized and holds the multicontact relay 90*CZ* up. This holds open the shunt around current element contact *CR* in the circuit to contactor *NR* and allows this current element contact to be effective. Likewise, a contact of 90*CZ* is held closed to make the current element lowering contact effective. The arrangement of the current element contacts is such that the demand for a raise in excitation is blocked if the current element finds that full load current is being exceeded. If current is high enough to demand reduction, the current element contact *CL* operates the motor-operated rheostat to lower the excitation.

During faults, the instantaneous overcurrent relay 90*CY* picks up, de-energizing relay 90*CX*, causing it to drop out. When 90*CX*, which is a magnetically damped induction-disk voltage relay, returns to its reset position, it closes its back contact and picks up relay 90*CZ* to put the current regulator in service again. Thus no attempt is made to regulate current or interfere with the operation of the voltage regulator during fault unless the fault remains on for an extended time.

The lockout and annunciator scheme is shown in Figure 10. The faults demanding lockout operate contacts that energize device 86. This is a manually reset relay which, when electrically tripped, closes its contacts and maintains them closed until it is reset. Its contacts drop out relay 4, energize the trip circuits of the various breakers, and set up the alarm circuit. Those conditions which require annunciation only, operate relay 30*X*, which is also a manually re-

set device. Like relay 86, it cannot be reset successfully unless the trouble contacts themselves have been reset. The contacts of 30*X* operate the bell alarm and lamp. In this circuit, the bell and the lamp will be cut off when 86 and 30*X* are reset. During the period when trouble is being located, the bell alarm may be cut off by picking up relay 30*Y* with a bell cutoff push-button, leaving the lamp lighted. Subsequent resetting of 86 or 30*X* will put out the lamp and drop out 30*Y* to re-establish the bell alarm circuit. There is also the provision that the bell will ring if it has been cut off with the push-button following the operation of one relay, say 86, and the other operates, in this case 30*X*. A break-contact of 30*X* in the coil circuit of relay 30*Y* is adjusted to open just as it begins to pick up, and it drops out relay 30*Y*. After relay 30*X* closes its make-contact, relay 30*Y* may be picked up again by the bell-cutoff push-button.

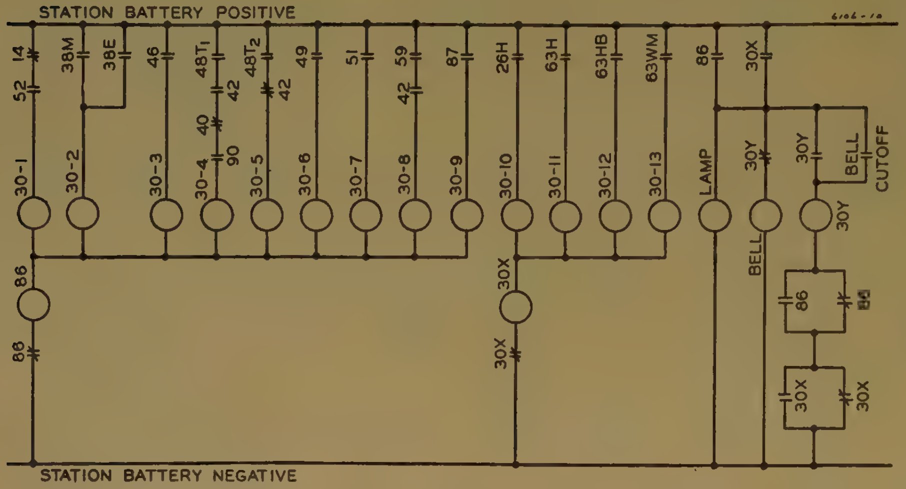


Figure 10. Annunciator schematic diagram showing alarm circuits and lockout relay

The circuit of the locked-rotor protective relay is shown in detail in Figure 11. The voltage appearing across the field-discharge resistor is compared with the voltage of the bus in such a way that any beat frequency between the two will operate relay 14*Z*. At standstill, a voltage is obtained from the field-discharge resistor that is equal in frequency and magnitude to the voltage received from the bus potential transformer. It may have any phase relation with this potential, however, and may or may not pick up relay 14*Z*. As the rotor begins to turn, the phase relations change and 14*Z* alternately is picked up and dropped out. This action is employed as an indication of rotation. If relay 14*Z* picks up, drops out, and picks up again, the relay chain consisting of 14*Y*1, 14*Y*2, and 14*Y*3 will energize relay 14*X*1, which in turn will hold relay 14 energized and prevent tripout through its back contact. Initially, relay 14*X*1, which is a pendulum relay, is picked up in the period of time between the closing of the starting breaker and the line breaker, and during the period when the rotor should start rotating, it is vibrating, alternately opening and closing its contacts in the circuit to the coil of relay 14. This relay is a slow-to-drop-out relay that will remain picked up even though 14*X*1 makes only intermittent contact. Should relay 14*Y*3 fail to energize relay 14*X*1 before it drops out, relay 14 will set up the circuit to the lockout relay to shut down the starting sequence.

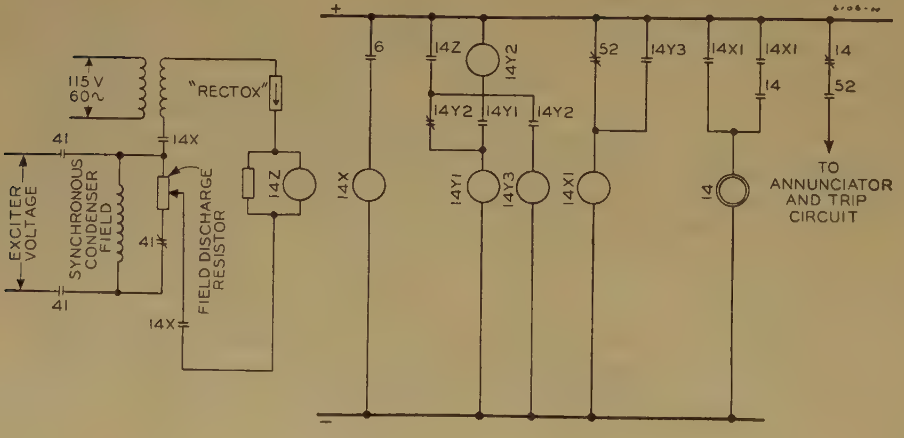


Figure 11. Locked rotor protective relay connection diagram

The stopping sequence begins with the de-energization of relay 4. Back contacts of relay 4 trip the line breaker, and if the starting sequence is incomplete, they will trip the starting breaker. When the line breaker comes out, it trips the running breaker, and the running breaker trips the field breaker. The field breaker shuts down the exciter by tripping the exciter motor breaker. The voltage regulator runs to the start position after the running breaker opens. The cooling-water valve motor runs to the close position when relay 4 drops out.

Emergency stopping does not respect sequential tripping. All breakers are tripped simultaneously by 86 and relay 4 is de-energized.

Conclusions

The operating complications involved in placing a synchronous condenser on a line demand that a certain portion of the control sequence be automatic. It is of practical advantage to design the sequence so that it is entirely automatic, allowing the user the choice of manual push-button control, remote supervisory control, or full automatic control.

Protective devices can be provided to protect against any contingency, and these devices either may sound an alarm to indicate an abnormal condition requiring attention or can lockout the machine in the presence of a dangerous condition.

Electronic Register Control for Multicolor Printing

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Synopsis: The proper register for multicolor printing, that is, the printing of as many as five colors on top of one another to the required accuracy of a few thousandths of an inch, has been successfully solved by the use of phototubes and associated electronic equipment. In the system described, the electronic portion extends all the way from the phototubes to the thyratrons which supply the required current to energize the correcting motor armature in the proper direction and at the speed for correct register. A typical installation is shown in Figure 1.

The Problem

THE NUMBER of copies of each issue of present day popular magazines runs into the millions, and the only practical way to print such a large number of copies is by means of a continuous web press. This means that the colored illustrations must be printed on large rotogravure presses in which the paper speed may be as high as 1,000 feet per minute. At this high speed each color, in turn, must be printed, then dried while in motion so that it will not smudge as the next color is printed. This wetting and drying of the paper causes shrinkage and stretching which differs with each individual roll of paper, depending on its composition and the amount of latent moisture present. The paper often moves 30 to 40 feet through the drying section before it reaches the next impression cylinder. Also if a printing cylinder is even a few thousandths of an inch over- or under-size, the tension between stands will differ greatly. A typical press and control are shown schematically in Figure 2.

A particularly critical situation occurs when a "flying paste" is made from the end of one roll to the beginning of the next. As the splice passes through each succeeding section of the press, the tension drops to a low value, and the critical tension balance necessary for correct register is completely lost. If no correction is made, as many as 300 copies might be spoiled before the correct tension balance is regained.

Before the application of electronic control, the most usual way to check register was by means of a battery of control handwheels located at the folder. At intervals the pressman would draw a cut-and-folded copy from the outcoming stream and unfold it to be scanned for evidence of misregister. He then would make the proper corrections by turning the appropriate handwheel the amount necessary as judged from long experience. Even though the pressman was continually in motion as many as 50 copies might be spoiled before the correction could be made.

Needless to say, holding register manually has never been a pleasant job, particularly if the paper has not been permitted to season, that is, to remain in a special storeroom long enough for every roll to attain the same conditions of temperature and humidity. No attempt was made to hold register through a splice, and it was usually assumed that all copies printed during acceleration or deceleration of the press would be spoiled.

Fundamentals of the Electronic System

The first fundamental requirement of the electronic control is to check each printing cylinder after the first to find if it is in the correct position to lay its impression exactly upon the previously printed one. Next, when an error has been found, means must be set in motion to correct the error. Since the accuracy of register must be within 0.003 inch to be acceptable, it will be seen that the system must respond to signals as short as 0.00001 of a second, or ten microseconds, if satisfactory operation is to be obtained. A complete electronic system is the only practical solution.

The second requirement is to apply power to the correcting motor, as indicated by the amount of error found. Although other means could be used to accomplish this function, electronic circuits using thyratrons eliminated noisy moving parts which, at the large number of operations required, would present a serious maintenance problem.

The Error Detecting Equipment

In order to detect the error, we must have a standard indication of the position of the paper at any instant and the position of the cylinder at that same instant. Along a margin of the paper where it can be trimmed off, or placed in the fold, are printed a series of register marks in the first color to be used. This is usually yellow. Typical marks are 0.020 inch in the direction of paper motion, and about one half inch across. Although a mark only one fourth inch long is needed, the extra length permits greater latitude in lining of the paper and in correcting side lay or side register. These marks are equally spaced, as seen in Figure 3, usually 16 to the cylinder circumference. While fewer marks can be used, the larger number permits greater correction speed and thus higher accuracy and more nearly perfect copies.

As this thin yellow mark arrives near the succeeding cylinders for printing of

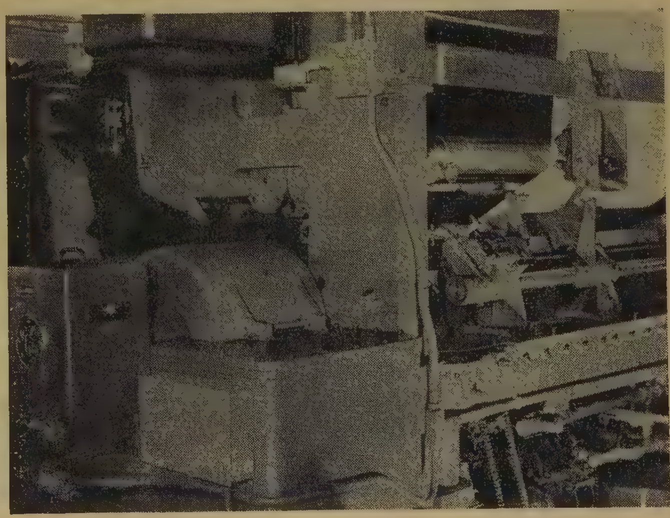

Figure 1. At each printing cylinder the web scanning head at right and the cylinder scanning head at lower left feed signals to the mixing panel at top. This determines proper operation of the correcting motor at lower right

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the other colors, its presence is detected by a phototube in the web head through an optical system which produces the effect of a very narrow slit, only a few thousandths of an inch wide. A dark blue glass filter provides a much better contrast between the yellow mark and the white paper.

The indication of the passage of the mark in front of the phototube leaves the web head in the form of a square wave of voltage of very short duration.

Detecting the Cylinder Position

Likewise, the position of the printing cylinder must be found so it can be checked against the relative position of the web. The basic indicating element is a disk or drum which contains the same number of axial slits as there are register marks on the first printing cylinder. A typical disk is ten inches in diameter, and the slits are one fourth inch wide. They are laid out so that the trailing edges of the slits are spaced very accurately around the circumference, as seen in Figure 4. The indicating point is not the whole slit but the trailing edge. An optical system directs light through this slit to a stationary slit before a phototube. The phototube and optical system can be moved around the drum to allow for setting up and for adjustment. The signal sent from this cylinder scanning head is a square voltage wave much broader than that from the web head, but which has a very accurate and abrupt trailing edge.

The Mixing Panel

The mixing panel, mounted on the press near each cylinder, may be considered the nerve center of the register equipment. Here the signals from the web mark and the cylinder disk slits are checked against each other and the proper correction action is initiated. For correct register, the cylinder disk phototube is moved until the trailing edge voltage change splits exactly the signal received from the 0.020 inch web mark. When this occurs, the correction circuit is balanced and no corrective action takes place. However, should the edge arrive even a few microseconds earlier or later than the midpoint of the mark, the difference is detected in the mixing panel, and the proper correction action is started. This comparison is made by charging one small capacitor as a result of the presence of the mark signal before the edge arrives, and charging a second capacitor after its arrival. These two charges then are compared and the difference is stretched out in time and amplified to apply the necessary correction. This correction signal, in the form of a small current of a few milliamperes, leaves the mixing panel in one of two leads, depending on which direction the correction is to be made. Perhaps the schematic drawing of Figure 5 will assist in clarifying the mixing panel action.

Since at this point the correcting motor control signal is at a low power level and in a low impedance circuit, it provides an ideal place to insert the push button stations desired for manual control of the correcting motor during setting up.

The mixing panel also contains an indicator "magic-eye" tube to assist the operator in setting up the panel and obtaining the original register adjustment.

The Motor Power Panel

Now that the correction signal has been obtained, it remains only to amplify the power for the correction, that is, the energy for the correcting motor. These correcting motors are usually d-c shunt gear-motors of about one half horsepower rating. They may operate differentials to change the relative position of the printing cylinders or may move the usual compensating rolls up and down.

The motors are run with full field strength to obtain maximum torque and fastest correction. The motor fields usually are fed from a common rectifier using phanotron tubes. The motor armatures are fed from 60-cycle alternating current through a pair of inverse-parallel thyratrons. Since the thyratrons are controlled rectifiers, if the first thyratron conducts a pulsating direct current, current flows through the motor armature in one direction. However, if the second thyratron conducts instead, the current flows through the armature in the opposite direction and opposite rotation is obtained. An iron core reactor in series with the motor armature smooths out the current and prevents excessive armature heating.

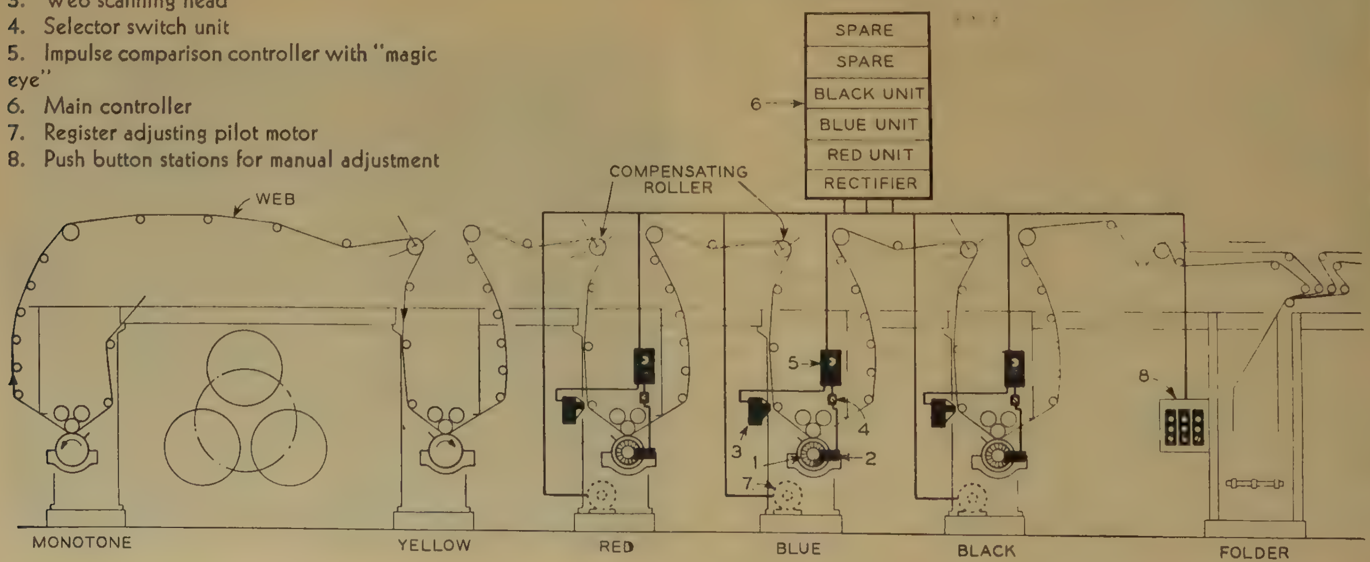
The only thing now required is a connecting link between the correcting signal from the mixing panel and the thyratron grid. This is supplied in the form of small saturable reactors in a bridge circuit which changes the phase of the grid. These reactors, of course, are saturated by the direct current from the mixing panel to regulate the grid phase as required. Figure 6 shows a complete motor control cabinet.

The Electric Circuit

Now that we have obtained a bird's-eye view of the operation of the equipment, let us retrace our steps and see the electric circuit used to obtain these results.

Figure 2. Schematic drawing of a multicolor gravure press with photoelectric register control

1. Cylinder register scanning disk
2. Cylinder scanning head
3. Web scanning head
4. Selector switch unit
5. Impulse comparison controller with "magic eye"
6. Main controller
7. Register adjusting pilot motor
8. Push button stations for manual adjustment



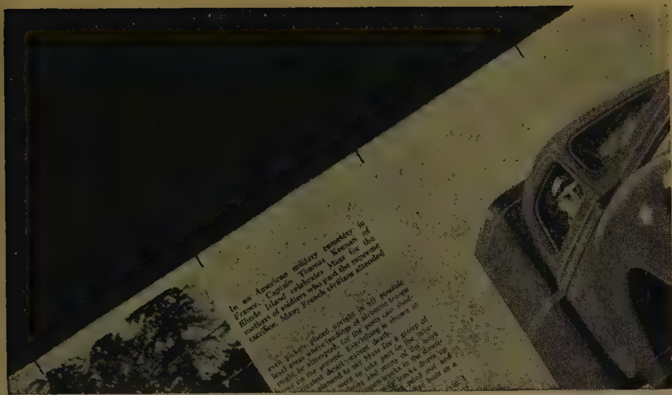


Figure 3. The register marks along the edge of the sheet usually are printed yellow and are almost invisible to the eye

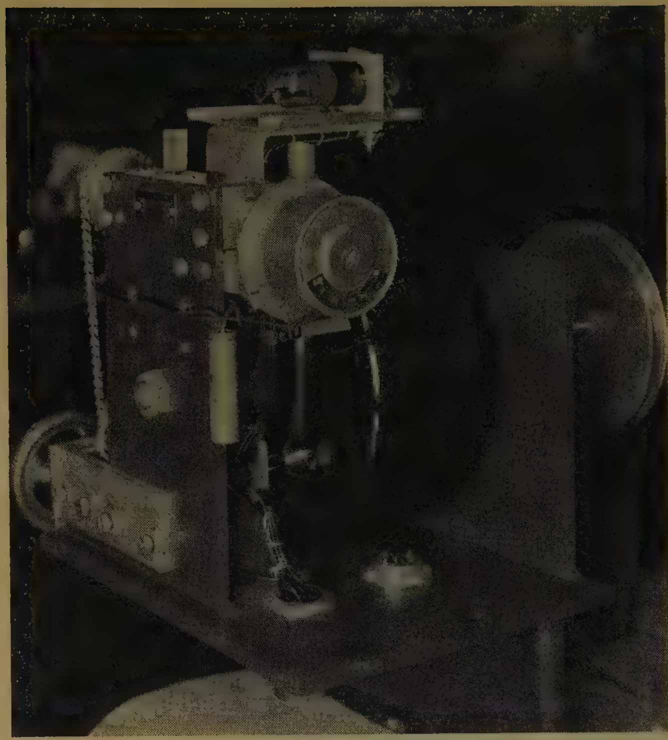


Figure 4. Cylinder scanning head showing slits in disk, lamp in center and phototube at top

Vernier adjustment motor is at upper right. (View taken during test with covers removed)

The Web Scanning Head Circuit

The web head consists of an automobile headlamp-type bulb and an optical system to flood the path of the register mark with an intense spot of light at the point where the phototubes must scan it. An optical system picks up the light from the mark and conducts it to the slit in front of the phototube, and the blue glass filter, mentioned previously, provides a more positive light change.

Because the change in light on the phototube as the mark passes is so extremely small and so rapid, it is necessary to use an extremely sensitive amplifier to provide a signal sufficiently strong to be transmitted to the mixing panel. A two-stage amplifier is used, and if the signal is to be transmitted for a considerable distance, a third tube as a cathode-follower is included. The tube chassis must be shock mounted to prevent the vibration of the press from causing bad microphonic disturbances. Figure 7 shows the chassis removed from the web head.

Although there are no arcing parts in the head, it always is made dustproof and when conditions require, explosion-proof.

Cylinder Scanning Head

Since the slits in the cylinder reference disks or drums permit operation on direct light transmission rather than by reflected light, a single stage of amplification is sufficient to provide a satisfactory signal of the passage of the slit. Here again, however, if the signal must be transmitted a long distance, the cathode-follower tube should be added.

A special feature of the cylinder scanning head is the mounting of the optical system and amplifier chassis. These are mounted on a radial arm which is pivoted at the cylinder axis so that the complete system can be swung around the disk in a limited arc. This is provided to obtain proper register in setting up the equipment and for the vernier control of register during operation. The arm is moved by a small reversible adjusting motor which is controlled by push buttons at the folder.

Although only about one half inch of each drum slit is required for operation,

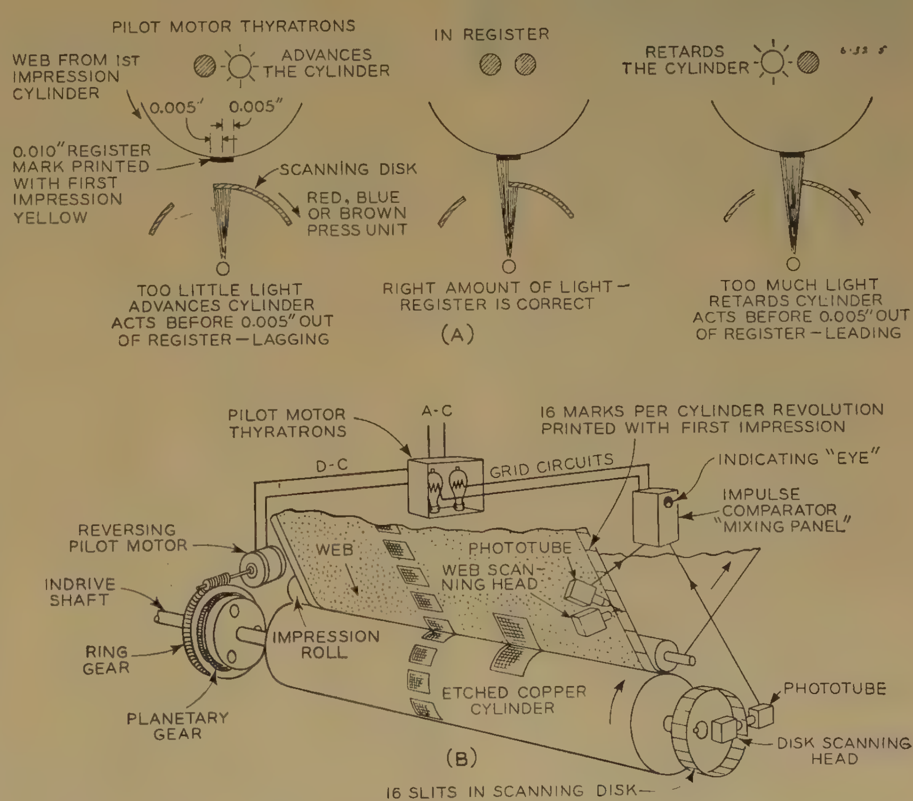


Figure 5. Schematic drawing to show the principle used to check register and to indicate the correction to be made

A. As impulse comparator evaluates the combined impulses from two phototubes
B. Photoelectric register control on four color press—one of four or more units shown

Figure 6. Motor control cabinet

There are three motor control panels at center, a spare panel at top, and the two rectifiers on the lowest panel

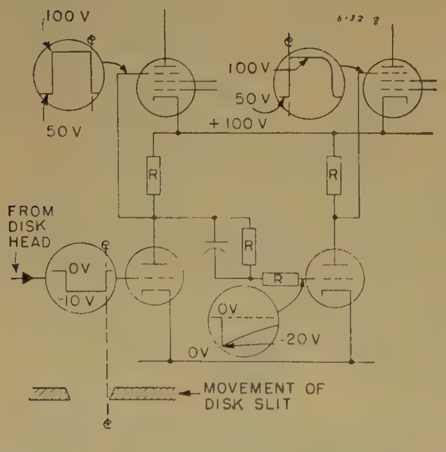


Figure 8. The action of the disk signal amplifier and phase-inverter

The wave shapes of the incoming and outgoing signals are shown also

e represents the instant that the trailing slit edge passes

Inserts are voltage wave forms

it is made much longer to permit the cylinder to be adjusted along its axis for best side-lay or register. This side register adjustment is usually a manual operation.

In some press installations a master cylinder scanning head is used rather than one at each printing cylinder. In this case the master unit is mounted on the main drive shaft, and optical systems for each color are located around it.

A single master cylinder scanning head permits a more compact installation than the separate heads, particularly when explosion-proof equipment must be furnished, but it can be used only with the compensating-roll type of correction.

Mixing Panel Circuits

The mixing panel must perform a number of functions. Let us take them up one at a time.

The incoming signal from the disk or cylinder scanning head is in the form of a square voltage wave which corresponds to the passage of the slit. In order to minimize the number of false operations due to random ink spots or paper blemishes, the region of active operation is limited to a fraction of an inch on each side of the correct register point. The signal from the slit provides a convenient live zone ahead of the correct point, while a similar active period after the slit has passed is obtained by means of a simple multivibrator circuit which uses grid rectification to charge a small capacitor during the slit passage to stop conduction in the trailing tube for a corresponding period. This twin triode also acts as a phase inverter. That is, one anode ceases to conduct during the time of the first active period, and then the second anode ceases conduction during the second active period. Figure 8 shows this circuit in simplified form.

The next step is that of comparing the disk head signal with that from the web head. This is done by means of the number one grid and number three grid of a pentode tube. Since the grids of a pentode act as valves in series, if either grid is sufficiently negative with respect to the cathode, there will be no current flow in the tube. The action is as follows: Two pentode tubes are used. The number one grid of both tubes normally is held negative, but rises to cathode potential when the marks on the web pass the phototube. But no current can flow unless the number three (suppressor) grid is also at cathode potential. These grids are controlled by the output of the two anodes from the twin-triode phase inverter, which cease conduction respectively before and after the register point. Thus,

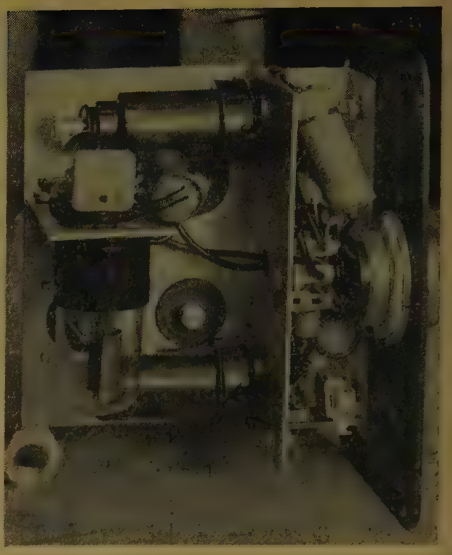


Figure 7. The web scanning head chassis

The small change in phototube current due to the passing register mark is amplified by the two pentodes at the right. The tube panel is isolated from press vibration by the three rubber mountings seen behind the tubes

one pentode conducts when the mark is passing and the slit is passing at the same time. The second pentode conducts if the mark passes for a short time after the slit has passed, as determined by the timing circuit. If the mark passes exactly in register, the first pentode will conduct while the first half of the mark is passing, and the second half passing after the slit will permit the second pentode to conduct.

If the mark does not appear, there will be no conduction in either tube and no correction made. This is important, for it means that if the mark does not print, the correcting action ceases, but there will be no false correction as in some systems.

The mixing circuit and typical wave forms are shown in Figure 9. Both the twin-triode and the pentodes are operated from cutoff to saturation and thus act as limiters to produce uniform voltage

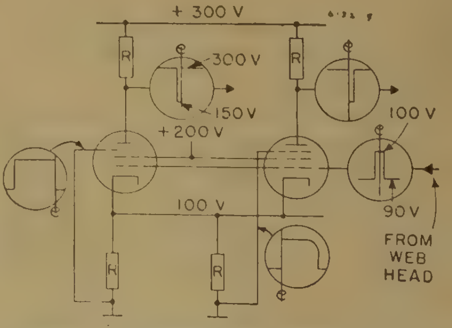


Figure 9. The circuit and wave forms of the pentode-tube mixing circuit

changes at the output. From the output of the pentode we now have signals which indicate the direction required for the correction, and by their relative duration, indicate the amount of correction needed.

The final function of the mixing panel is that of stretching out the fast microsecond pulses of the pentode output to a time sufficient for operation of the correcting motor. This is done by means of small capacitors which store the energy presented to them at each mark passage and permit it to leak off at a rate slow enough to permit motor operation to take place. This pulse-stretching circuit is shown in Figure 10.

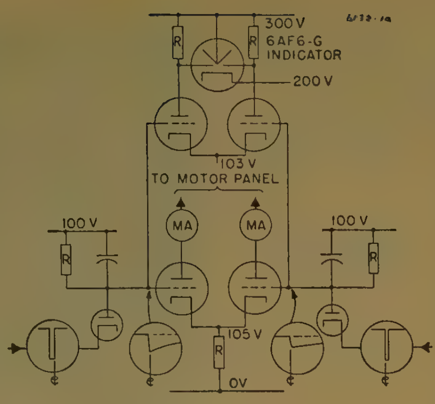


Figure 10. The correction signal pulse-stretching circuit, the output triode amplifiers, and the "magic-eye" indicator (6AF6-G) tube circuit

Assuming register mark center arrives before slit edge

Each time one of the two pentodes conducts, its corresponding capacitor begins to charge through the small rectifier tube. The capacitor charge next is amplified by a triode tube, and the output of this tube is passed on to the motor panel.

At this point, the action may be summarized thus. If the trailing edge of the disk slit passes at the same instant as the midpoint of the register mark,* equal direct current will flow from the two triodes at the output of the mixing panel. If the register mark passes somewhat early so that there is an excess of mark before the slit edge passes, there will be a larger current in one triode, or if the register mark arrives late, there will be a larger current in the second triode output.

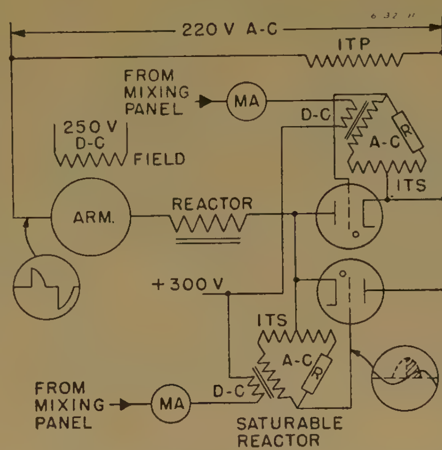


Figure 11. The power circuit and the grid control circuit for the correcting motor

The motor field is supplied from the common rectifier

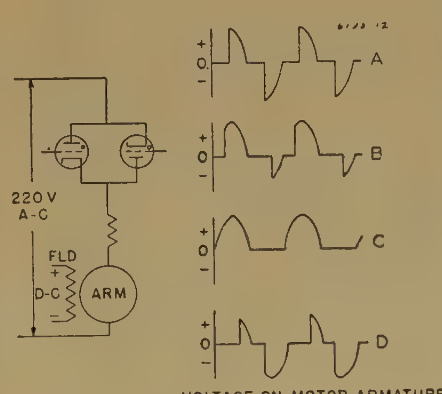


Figure 12. Representative wave forms of voltage applied to the correcting motor armature under various conditions of register error

A—Balanced, no motion
B—Forward slow
C—Forward fast
D—Reverse slow

The "Magic-Eye" Indicator

In addition to the control circuit, the mixing panel also has an auxiliary circuit, consisting of a twin triode and a double "magic-eye" tube which permits a visual indication of whether the register is balanced or is off in either direction. This circuit also is fed from the small pulse-stretching capacitors. If the mixing panel is mounted remote from the press, the indicator tubes may be mounted at the folder.

Manual Push Button Control of Correcting Motors

On its way to the motor control panel, the output of the mixing panel passes through the push button stations used for manual control of the correcting motor. These push buttons are used principally for setting up the work at the beginning of a run. When they are operated, the circuit from the mixing panel is broken, and direct current from a separate source is fed into the motor-panel circuit to operate the motor in the desired direction.

Motor Panels

For convenience, the motor panels usually are assembled together in a large floor-mounted cabinet. This cabinet also contains the common rectifier supply for the motor field and a rectifier to supply the mixing panels. A tap also is taken off this second rectifier supply for the push button operation. Each motor panel is complete in itself with jack and plug connections for all leads. In some installations, a spare motor panel is included in which the thyratrons are kept at operating temperature so that should a tube failure occur, the spare panel may be substituted for the inoperative circuit and a minimum of time lost.

The armature circuit consists of a reactor in series with the armature and the two inverse-parallel thyratrons which are used to produce the reversing action of the armature. Each thyratron has its own grid circuit which consists of a center-tapped transformer feeding a saturable reactor-resistor bridge, as shown in Figure 11. Since the a-c and d-c windings of the saturable reactors are electrically separate, the motor circuit is insulated from the control circuit coming from the mixing panel.

In operation, a larger current flowing in the d-c winding of the saturable reactor decreases its reactance, and the grid of the thyratron is phased forward. This permits the thyratron to conduct earlier in the cycle, and more current flows through the motor armature. The mixing panel circuit is so designed that, when register is correct, sufficient current will flow into the saturable reactors of each of the thyratron grid circuits to cause both thyratrons to conduct for approximately one-quarter cycle. The opposing currents balance in the motor armature so that it does not rotate but vibrates slightly at 60 cycles. However, when the register is out by even a few thousandths of an inch, the current to the reactors is unbalanced so that the thyratron currents are not balanced and the motor moves in the direction determined by the larger current. The shape of the voltage wave applied to the motor for various conditions is shown in Figure 12.

When the controlling current is completely unbalanced, that is, when the register marks arrive completely before or after the disk edge arrives, full correcting motor speed is obtained. This is reduced to zero as the slit edge intercept comes nearer to the center of the register mark. Antihunt circuits in the mixing panel prevent overshooting.

Aids for Maintenance and Repair

In the printing business it especially is important that production schedules be

Figure 13. Mixing panel with cover removed

This shows "magic-eye" indicator tube at top, milliammeter jacks at left-center, test terminals at center, and connecting leads plugged in at bottom

Figure 14. Preventative maintenance

Checking the equipment operation with a small cathode ray oscillograph

met. Therefore, even short delays in press operation are serious. It has already been mentioned that the motor panels are provided with plug and jack connections and that sometimes a spare motor panel is added and kept warmed up for instant substitution without the delay of even the five minutes needed to preheat the thyratron cathodes. Although there are no moving or wearing parts on the web and disk scanning heads, or on the mixing panel, all of these are also provided with plug and jack connections, and a spare of each is often kept on hand for substitution whenever needed.

To assist in preventative maintenance and trouble shooting, milliammeter jacks and test terminals are located on the mixing panel, as seen in Figure 13. The milliampere readings of the output to the motor panel permit a quick checkup of the over-all mixing panel operation, while the test terminals permit key potentials to be seen on a cathode-ray oscilloscope to check the operation of the two scanning heads and the internal circuits of the mixing panel. To the experienced operator the characteristic flicker of the "magic-eye" indicator tube can convey much information about the over-all operation of the equipment.

Putting the Equipment Into Operation

In conclusion, let us follow the press operator as he puts the register equipment into operation when starting a new run on the press. The inkwells are filled, the paper is threaded, and the backing rolls are let down. The first copies coming off the folder look very little like the finished product, but if the press crew is experienced, the register will be within a fraction of an inch correct on all colors. Standing at the folder with a copy in front of him, the head pressman then presses the proper push buttons operating the one half-horsepower correcting motors to bring the cylinders into approximate register with the first color, usually yellow. First the red impression from the second cylinder comes into line, next the blue, and finally the black from the last cylinder.

The press is then run up to a moderate speed so that the normal tension in the web is obtained, and the register is again checked and corrected by the push buttons. The register may now be assumed to be good enough to be placed on the automatic operation. To do this the pressman first operates on the red cylinder disk scanning head, and moves the scanning head arm until register is obtained as indicated by the "magic-eye" operation, indicating a balance of the mixing panel output. He then throws the switch to automatic, and the red is locked in step with the yellow. He then repeats this process on the blue and on the black cylinder. Finally he returns to his normal station near the folder and checks the outcoming copies. Probably during the time he has been away the register has crept off a few thousandths. He now brings this back into accurate register by again operating the small reversible adjustment motor on the disk head optical system. Operation of the push button controlling the correcting motor at this time will have no effect because the automatic operation immediately would bring the register back to the previous setting.

After these refinements have been made, the register is locked in step in the correct position and will hold this through wide changes in press speed, through splices, and through the many differences in paper which may be found, such as tension, humidity, surface conditions, and others.

Conclusion

It was said in years gone by that only one copy in 25 was perfectly registered. With electronic register control, fewer than one in 25 will be found which are not perfectly registered. Electronics has provided a quiet, precise method for holding register. Since there is no wear of critical parts the high accuracy which can be obtained will continue throughout years of untiring service.

TRANSACTIONS SECTION

Preprint of Corresponding Pages From the Current Annual AIEE Transactions Volume

Design of D-C Auxiliary Controllers for Marine Service

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MEMBER AIEE

CONSIDERABLE progress has been made in recent years in the design of marine control equipment. Under the impact of the war, the performance of controllers has been improved and better equipment of lighter weight has been made available. It is the purpose of this paper to discuss some of the more important design features on which the controller performance is based.

Controller Components

The function of a motor starter is to connect the motor across the line without excessive disturbance to the motor or the power supply. The simpler the equipment used for this purpose, the more satisfactory, in general, it will be in service.

There are a great variety of motor starters, ranging from the simple manual line starter to the more complex magnetic controller. Different methods of automatic acceleration are used with the number of starting steps varying to suit the application and rating of the motor. Other design features, such as overload protection and selection of resistors have an equally important bearing on the controller performance.

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Number of Starting Steps

The current surges during starting of a motor depend primarily upon the number of starting steps and the regulation of the motor. This relation can be solved mathematically but perhaps is more clearly illustrated in Figure 1 from a typical shunt motor with 10-per-cent speed regulation and a constant torque load of 100 per cent.

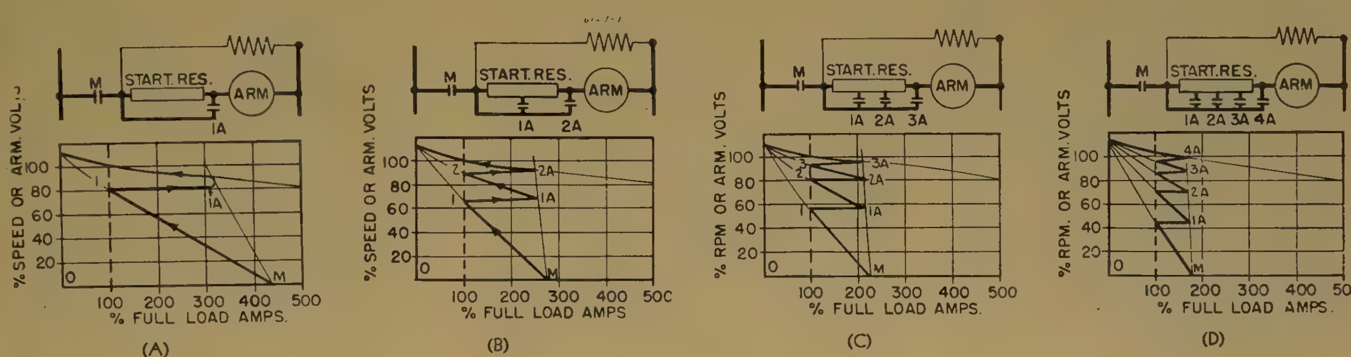


Figure 1. Schematic diagram and current surges for various numbers of starting steps

On closing the line contactor M, the armature current increases from 0 to M. As the motor accelerates, this current decreases to one, at which point the accelerating contactor $1A$ cuts out the starting resistor and connects the motor across the line. For multistep starters, this process is repeated by operating accelerators $1A$, $2A$, $3A$, $4A$, and so forth, in sequence until the starting sequence is completed.

Figure 1 illustrates clearly that the starting current surges can be reduced by increasing the number of steps. This fact is taken into consideration by providing more steps for large motors than for small motors. Table I shows standard number of starting steps and approximate maximum current surges for various motor ratings.

Further reference to Figure 1 indicates that the starting current surge is made higher for the low armature speeds than for the high armature speeds. This is done because the average motor can commutate current surges better at low speed than at high speed, a fact which is well illustrated in Figure 2.

This oscillogram shows a starting test with a standard 3-horsepower 250-volt motor connected directly across the line without a resistor step. Commutation was fairly good at low speed even though the current inrush was about 1,300 per cent. At approximately full speed, however, the motor flashed over, although the current had decreased to about 750 per cent.

Of additional interest is the fluctuation of the field current I_f during the starting of the motor. This current suddenly increased several hundred per cent, which indicates a pronounced weakening of the flux due to the armature reaction. The subsequent dip of I_f below the normal value is due to the line voltage disturbance caused by the short circuit.

Applications with high inertia loads

Table I. Number of Starting Steps and Maximum Current Surges for Various Motor Ratings

Hp at 230 V	Number of Starting Steps	Approximate Current Surge, Per Cent Full Load Amp
1/2	0	1,000
3 to 5	1	450
7 1/2 to 25	2	275
30 to 40	3	225
50 to 150	4	175

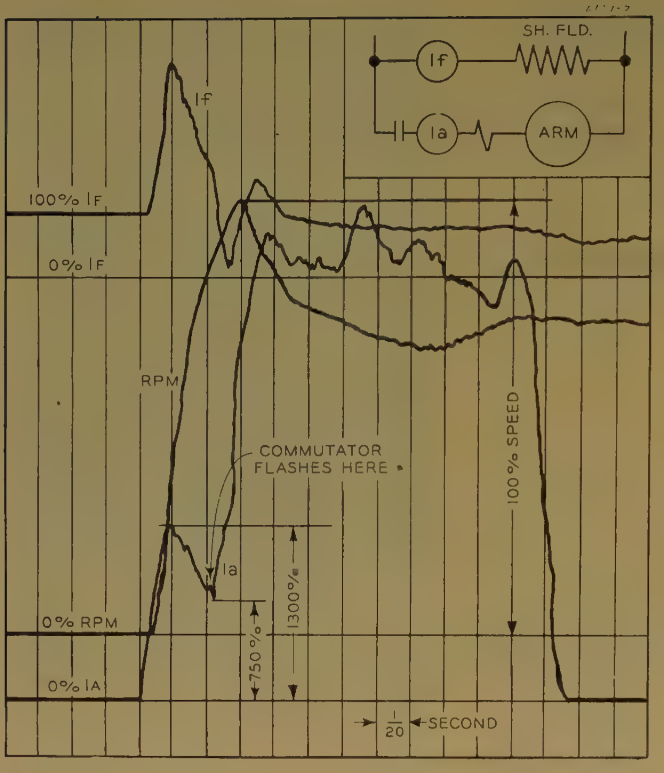


Figure 2 (left). Oscillogram showing line starting of 3-horsepower 250-volt shunt motor

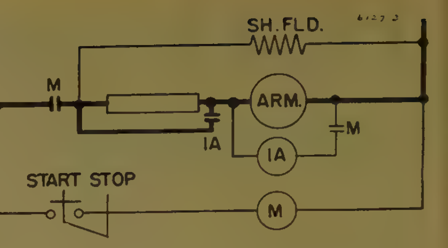


Figure 3. Starter diagram with counter-electromotive force acceleration

such as fans and oil purifiers are more critical and possibly should have more starting steps to keep down the current surges.

Methods of Starting of Motors

Various methods are available for starting of motors, namely—

(*a*). Manual starting.
(*b*). Counter-electromotive force starting.
(*c*). Current limit acceleration.
(*d*). Definite time acceleration.

Manual Starter. This type of starter undoubtedly provides the simplest means of starting a motor. It is particularly suitable for motors up to one half of one horsepower where the stalled armature current does not exceed 1,000 per cent. Sometimes it is provided with a starting step for use with motors up to five horsepower. In such cases, the operating lever may be equipped with a mechanical offset to make the operator pause on the start position before going to the run position.

In general, it is not desirable to handle large currents with manual controllers since there is no assurance of positive contact movement, which is essential for satisfactory operation.

Counter-Electromotive Force Starting. This method measures the armature voltage and cuts out the starting resistors when this voltage has reached a predetermined value. A schematic diagram, which is self-explanatory, is shown in Figure 3.

The counter-electromotive force starter is somewhat critical in its adjustment and can be used only for single-step starters. Figure 1A shows that the accelerating contactor 1*A* has to be adjusted to operate at 70–80 per cent in order to avoid an excessive current surge when it closes. With possible line voltage drops of 20 per cent this setting is rather critical, particularly if the pull-in voltage is increased due to heating of the magnet coil.

Multistep starters of the counter-electromotive force type would require even a higher setting of the last accelerating contactor as shown in Figure 1B, which makes use of the counter-electromotive force method for multistep starters impractical.

Current Limit Acceleration. This method is suitable equally for starting low and high inertia loads and will bring the motor up to speed in a minimum of time. A one-step starter of this type is shown in Figure 4.

The accelerator 1*A* is of the spring closed type and has a main winding $1A_M$ to close the magnet and a holding winding $1A_H$ measuring current. See Figure 5. Depressing the start button first will insert the starting resistor and then close the line contactor *M*. As the motor accelerates and the current decreases, the holding coil $1A_H$ will let go and close contact 1*A*, thereby connecting the motor across the line.

The accelerators are calibrated usually to operate at about 135 per cent current so as to make sure that they function when starting heavy loads. Thus the current surges will be somewhat higher than indicated in Figure 1.

Definite Time Acceleration. Definite time starting probably is the most practical method of bringing a motor up to speed. It includes one or more accelerators adjusted to function in predetermined time intervals. A single-step starter is shown in Figure 6.

The accelerator or "timetactor" operates on the basis of the well-known and reliable flux decay principle and is shown in sizes of 100 and 300 amperes in Figure

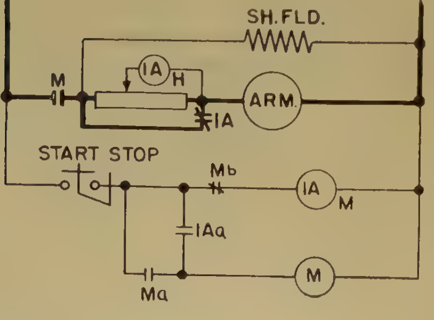


Figure 4 (left). Starter diagram with current limit acceleration

Figure 5 (right). Accelerating contactors for current limit or definite time acceleration

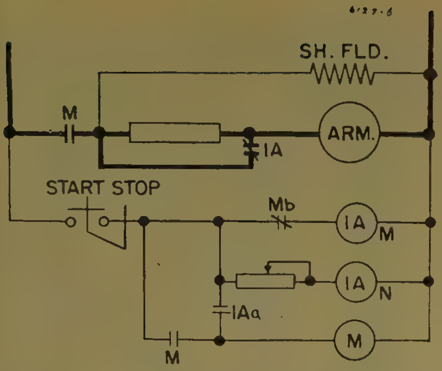


Figure 6. Starter diagram with definite time acceleration

5. In general, the time starter is identical to the controller of Figure 4 except that the accelerator is connected to obtain a fixed time instead of measuring current. Time adjustment is obtained by changing the current in the neutralizing winding $1A_M$.

There is, of course, the problem of selecting the proper time adjustment for the timetactors, so that the motor can accelerate without excessive current surges. This, however, should not be too difficult a task since starting conditions for the various ships' auxiliaries are well known and it is therefore a simple matter to adjust the accelerating time slightly longer than needed to obtain minimum current surges. Definite time starters can be tested and adjusted very readily with the motor disconnected and therefore have a decided advantage from a maintenance point of view.

Speed Control

The speed of a d-c motor may be controlled by varying the armature voltage or the strength of its field. Motors below five horsepower used for hull ventilation sometimes are provided with armature speed control in order to reduce the motor size. This gain, however, is practically offset by the increased size of the controller which has to absorb the resistor loss as shown in Figure 7.

In most cases, the motor speed is changed by varying the shunt field. Motors with speed changes of two to one or less have sufficient starting torque to permit good starting with weak field. The torque tests shown in Figure 8 were made with a 20-horsepower 230-volt motor of 600/1,800 rpm and a speed regulation of 20 per cent at the high speed. It is interesting to note that the flux weakening due to armature reaction is practically independent of the speed setting as evidenced by the bending of the three curves at 300 per cent load.

Controllers for motors with speed ranges greater than two to one are provided with a field relay in order to avoid too long an accelerating time particularly when used with fans of high inertia. The field relay is usually of the fluttering type.

Overload Protection

Each motor is provided with overload and short circuit protection. In line with well-established practice the overload relay is made part of the controller, while short-circuit protection in form of feeder breakers is located on the distribution board.

Protection of Continuously Rated Motors. Overload protection of these motors is usually obtained by means of a thermal relay set at 115 to 125 per cent of full load. Its time constant should be long enough to permit starting of the motor without tripping even though the starting time is long and the current surges are high.

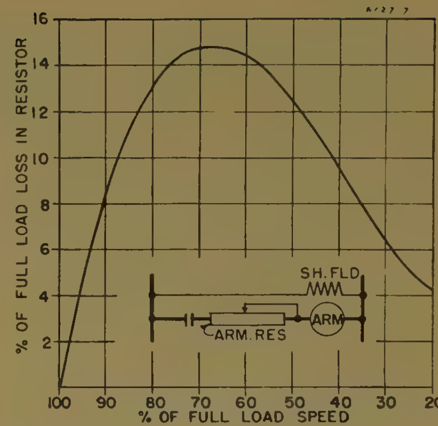


Figure 7. Resistor loss for armature speed control

In view of the high ambient temperatures encountered in engine rooms, the thermal overload relay is usually of the compensated type, that is, its calibration is made practically independent of the room temperature. Experience has shown that the noncompensated type is liable to trip under normal operation, thereby seriously interfering with the operation of the ship. A compensated overload relay is shown in Figure 9. It is provided with a special bimetal that corrects for changes in room temperature and thereby maintains the calibration practically constant.

Protection of Intermittently Rated Motors. There is a great deal of controversy as to whether overload protec-

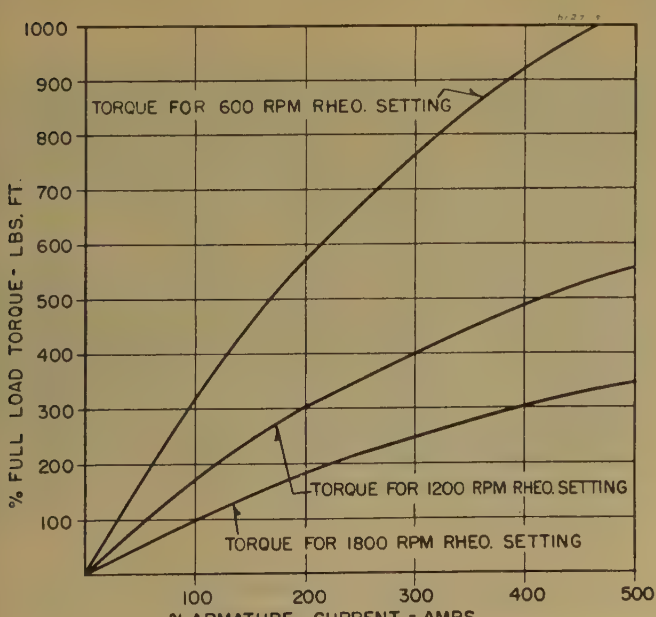


Figure 8. Starting torque of typical 20-horsepower 230-volt adjustable-speed shunt motor for different excitation and armature current

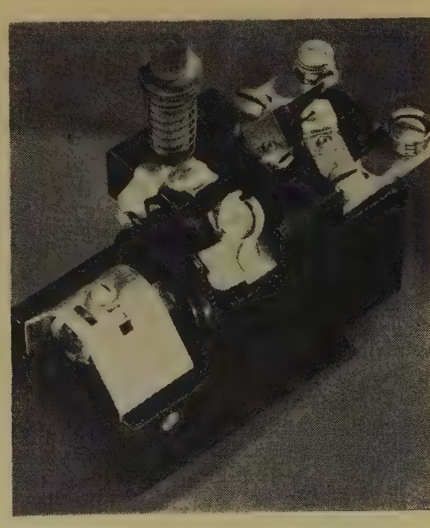


Figure 9. Compensated thermal overload relay for a-c and d-c applications

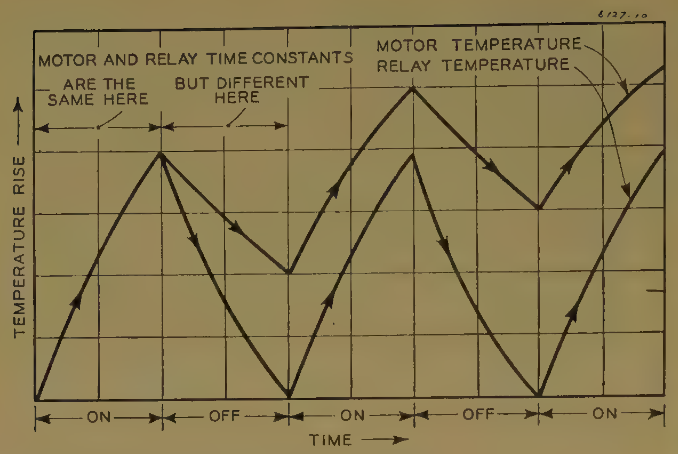


Figure 10. Heating curves showing inability of thermal overload relay to follow winding temperature of intermittently operating motor

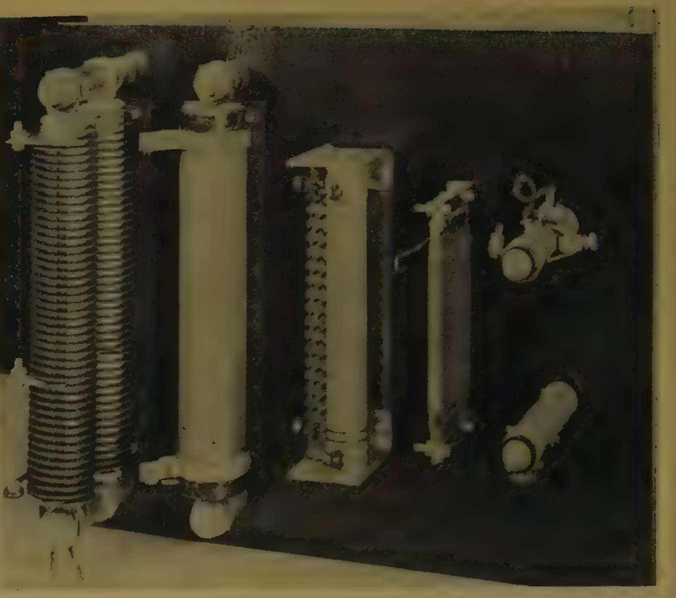


Figure 11. Various types of resistors

tion of intermittent motors by means of a thermal relay is practical. It is admitted that in order to be satisfactory such a relay should have the same time constant as the motor, that is, their temperatures should vary at the same rate when subject to various heating and cooling cycles.

This is a condition that is most difficult if not impossible to meet. Considering the fact that the time constant of motors may vary as much as three to one, depending upon their speed, ventilation, and size, it would not be feasible to design a multiplicity of relays to cover the whole range. Moreover, the co-ordinating of the time constants would be difficult since in most cases the desired information probably would not be available.

There is another important factor against matching time constants of motor and overload relay, namely, the unequal cooling of the motor while running and standing still. Tests show that with the motor standing still, the ventilation is very poor so that it will cool at a much slower rate than it heated during the running period.

Figure 10 shows that for intermittent operation the motor temperature will rise gradually above the relay temperature so that the motor is not protected even though the two time constants are matched while the motor is running.

For this reason, intermittently rated motors are protected by instantaneous overload relays set to trip at 200 to 300 per cent. They are intended to operate in case of emergency only, such as flashover of commutators or fouling of the anchor.

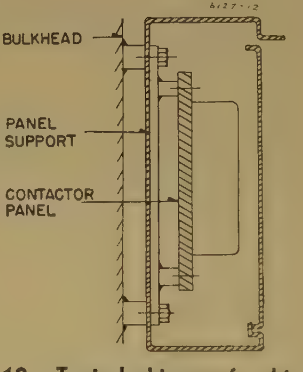


Figure 12. Typical drip-proof cabinet for bulkhead mounting

There is, however, the possibility of providing overload protection by means of thermoguards attached to the motor winding, thereby measuring motor temperature directly and stopping the motor when the temperature reaches a dangerous value.

Overload protection for capstans and windlasses is obtained with an instantane-

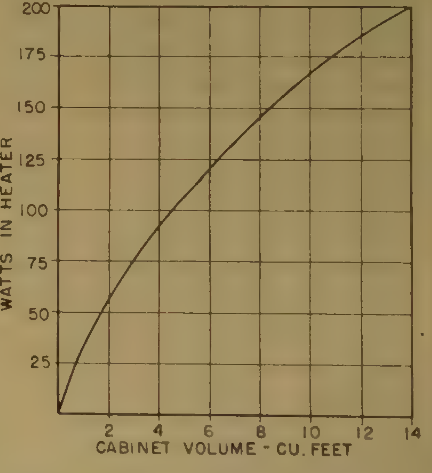


Figure 13. Heater capacity required for waterproof cabinets

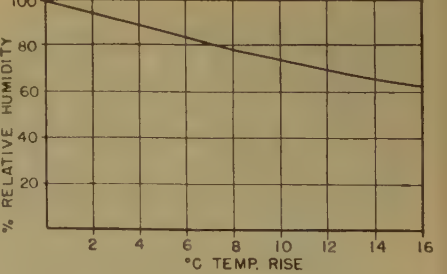


Figure 14. Relative humidity as a function of air temperature rise

Table II. Resistor Classification for Various Applications

Application	Duty	Resistor Class	Duty Cycle Sec On	Duty Cycle Sec Off
Starters below 40 hp	Starting	116	5	75
Starters above 40 hp	Starting	136	10	70
Windlass	Speed regulating	174DJ*	15	15
Capstan	Speed regulating	174J†	15	15
Cargo winch	Speed regulating	164D‡	15	30
Cable reel	Speed regulating	164	15	30
Boat hoist	Speed regulating	164	15	30

* Dynamic breaking and jamming protection with 125 per cent current for 5 minutes. Continuous running on last point lowering.

† Jamming protection with 125 per cent current for 5 minutes.

‡ Dynamic breaking in lowering direction.

The capacity of the resistors is based on good natural ventilation with not more than six tubes mounted on top of each other unless special baffles are provided.

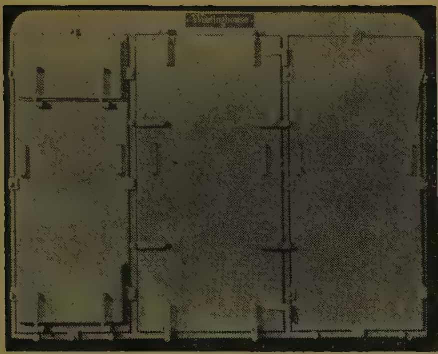


Figure 15. Typical waterproof cabinet for deck application

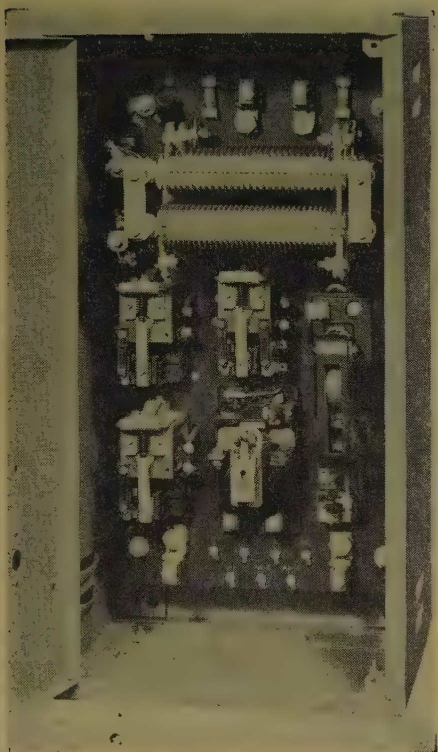


Figure 16. Typical drip-proof under-deck controller

ous jamming or step-back relay which inserts resistance in the armature circuit and reduces the stalled current to about 125 per cent rather than stopping the motor completely. This is done to maintain the tension in the anchor chain or the hawser while the ship is being warped or getting under way.

Resistors for Starting and Speed Regulating Duty

Practically without exception these resistors are of the nonbreakable and noncorrosive ribbon type using a nickel or chromium steel alloy suitable for 500–600 degrees centigrade continuous duty.

The maximum permissible temperature rise of a resistor depends a great deal on its location. When mounted in a controller cabinet, it is not desirable to exceed 365 degrees centigrade rise because it is close to apparatus with organic insulation.

However, it is satisfactory to go as high as 500 degrees centigrade rise when the resistors are mounted separately and are spaced sufficiently from other equipment to avoid undesirable heat transfer.

The tap or terminal is an important design detail of the resistors. Depending upon the design of the resistor, the terminal is either clamped, silver-soldered, or welded to the resistor ribbon. Any of these methods is satisfactory, although the welded terminal may be preferable, particularly for higher temperatures.

In order to insure a stable and permanent resistor connection, the latter must stay tight. Only in this way is it possible to keep the oxides away from the points of contact and maintain a metal-to-metal contact.

Two factors have to be considered when designing resistors for starting or regulating duty, namely, resistance and capacity. The selection of starting resistors is relatively simple since both factors are given by the number of steps and the accelerating time. Resistors for speed regulating duty, however, are more difficult to determine since the duty cycle depends a great deal on the judgment or habit of the operator.

Based on operating experience, resistor designs have been established for various applications, as shown in Table II. Special controllers with long accelerating time or heavy-duty cycle require, of course, more resistor capacity.

Controller Enclosures

Practically without exception, controller cabinets are built of structural steel in order to secure maximum strength and minimum weight at low cost. This applies to drip-proof as well as waterproof enclosures.

Designs for drip-proof cabinets are fairly well standardized. The contactor panel is mounted on rigid panel supports which, in turn, are bolted directly to the bulkhead as shown in Figure 12. This forms a strong mounting arrangement and permits a relatively light enclosure. Floor mounted drip-proof cabinets consist of the usual angle iron frame with removable sides and back to permit ready access to the apparatus. Ventilation of cabinets is usually obtained by means of louvers arranged in the side of the cabinet. This limited ventilation is sufficient for small losses such as caused by magnet coils. Larger losses, such as produced by speed regulating resistors,

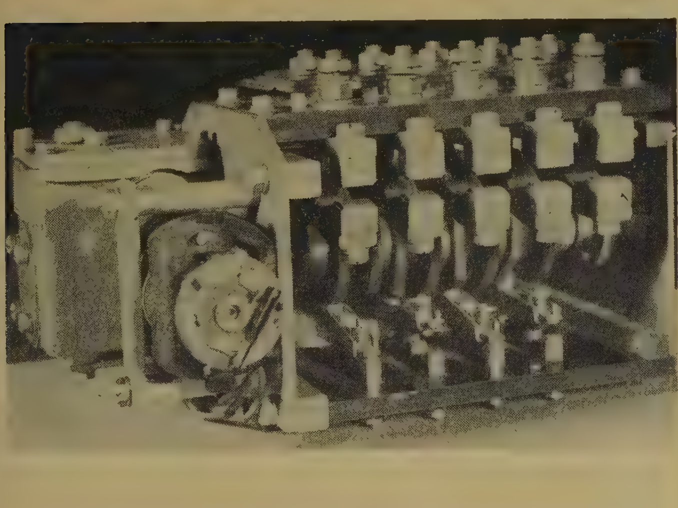


Figure 17. Accelerating drum for large motor starter with many steps

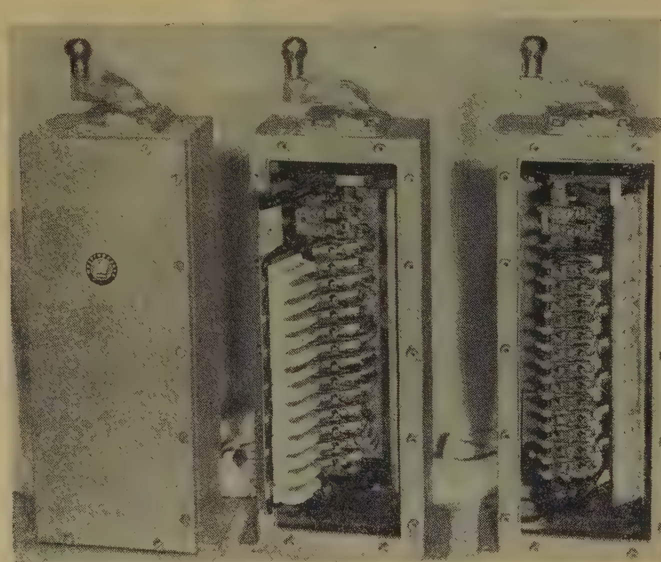
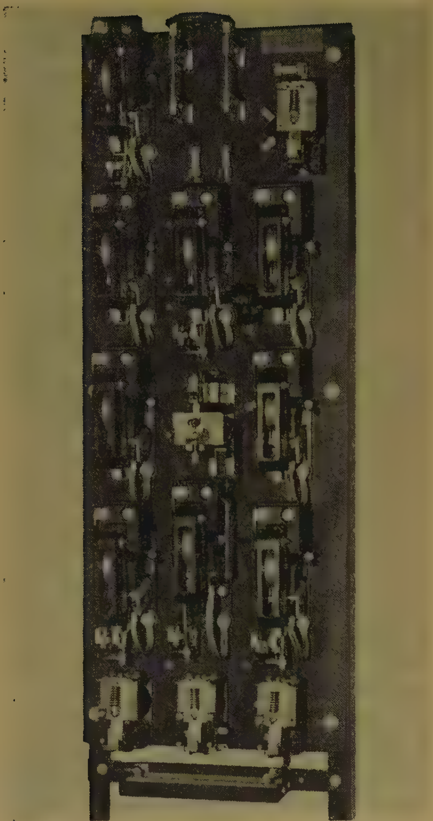
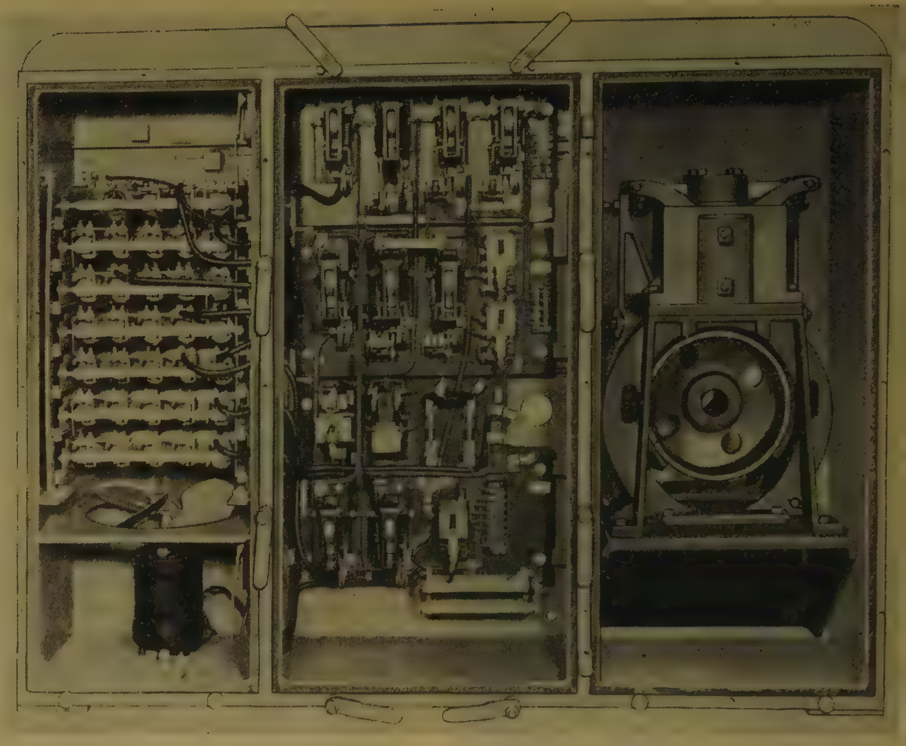


Figure 18. Waterproof manual controllers for deck auxiliaries



(A)

(B)



(C)

Figure 19. Magnetic controllers for deck auxiliaries

A. Open type
B. Drip-proof type
C. Waterproof type

require enclosures with raised top to permit better circulation of air.

Waterproof cabinets for deck mounting have to be made strong enough to withstand the impact of green water and to guard against corrosion eating through the walls in a relatively short time.

Corrosion of deck controllers is difficult to prevent because of condensation resulting from breathing action. The highly saturated air inside the cabinet is heated up to 50–60 degrees centigrade during a hot day and therefore, can hold considerable moisture. During the night this air is cooled off, and thus is unable to hold the water which is deposited on the inside surface, and finally collects at the bottom of the cabinet. This process is going on day after day so that under unfavorable conditions a sizable amount of water may collect in the cabinet, rendering the air so damp that even the best finish will break down after a while. There are two effective means of minimizing this corrosion. One consists in providing heaters inside the cabinet to keep the air always above the dew point. The watts required for this purpose can be determined from Figure 13. Experience has shown that a temperature rise of the air inside the cabinet of about 10 degrees centigrade is sufficient to keep the cabinet dry. The reduction of relative humidity by heating the air is illustrated in Figure 14.

An additional improvement is obtained by providing a check valve at the bottom of the cabinet to drain any possible water and to provide some circulation of fresh air. Actual tests in the field have shown that a considerable improvement is obtained with surprisingly small openings.

It cannot be emphasized too much to provide substantial bolts not less than 3/8 inch diameter for mounting of doors or covers. Sizes below that will be twisted off when trying to force a corroded thread. Doors that have to be opened very readily sometimes are equipped with locking dogs as illustrated in Figure 15. They have to be made substantial enough to withstand rough handling when covered with ice.

Controller Assemblies

It is not the purpose of this paper to give a detailed description of the various applications. Rather than do this it is proposed to show some typical controllers which illustrate the lastest design trends.

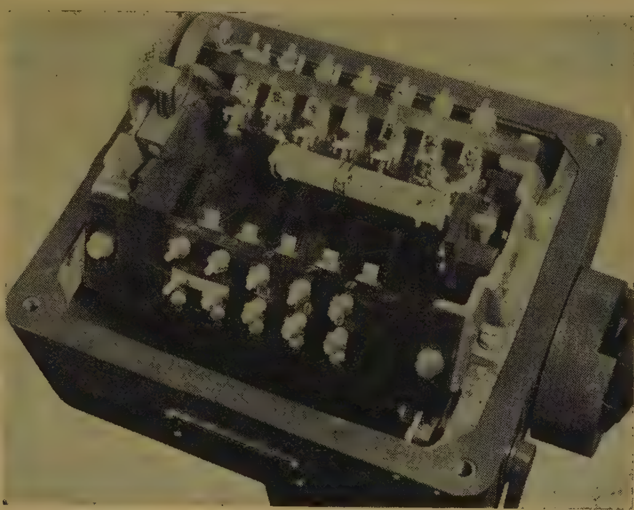
Underdeck Controllers

Figure 16 shows a typical three-step starter for a 25-horsepower motor. All controller components are mounted in front of the panel to permit ready inspection and removal of parts if necessary. Studs with solderless lugs are arranged

top and bottom for external cables, the wiring of which is facilitated by detachable and standardized lead plates of ample size. Whenever possible, the push button is mounted on the starter door to obtain a unit assembly and to reduce wiring at the shipyard to a minimum.

Large starters of high-current capacity requiring many starting steps very often are provided with a motor-operated accelerating drum as per Figure 17, which combines one line contactor and all accelerators in one single and compact unit. The spring-closed cam-opened switch units are provided with silver tips and cut out the starting resistor in predetermined intervals. The speed of the pilot motor is adjustable within a one to two speed range.

Figure 20. Two views of typical waterproof master switch for deck controller



CONTROLLERS FOR DECK AUXILIARIES

Manual as well as magnetic controllers are provided for winches, capstans, and windlasses. Manual controllers as shown in Figure 18 are limited usually to smaller motors not exceeding 30 horsepower at 230 volts or 15 horsepower at 115 volts, because experience has shown that maintenance is excessive for current ratings above these values.

Magnetic controllers are being supplied with open, drip-proof, or waterproof unit construction as shown in Figures 19A, 19B, and 19C. There has been a great tendency in recent years to favor the unit construction, particularly for cargo winches, since it reduced the installation work on the part of the shipyard to a very minimum. Its possible weakness is maintenance at sea, particularly in bad weather, although experience so far has been very satisfactory.

Master switches are invariably of the waterproof construction, either cast or sheet steel and with an offset in the handle movement to provide a positive stop in the off position. A safe-on auxiliary switch is also provided to permit shutting down the control circuit when the equipment is not in operation. See Figure 20.

Very often it is preferred to make the master switch spring returned to obtain the so-called "dead man's feature" and to make sure that the motor cannot stay energized and accidentally burn out while the operator has "gone out for lunch." A master switch of this type is illustrated in Figure 21. Its lever is arranged so that the switch can be operated conveniently either with the stevedore standing beside the controller or sitting on it. This latter feature is desirable particularly for cargo handling where operation extends over many hours.

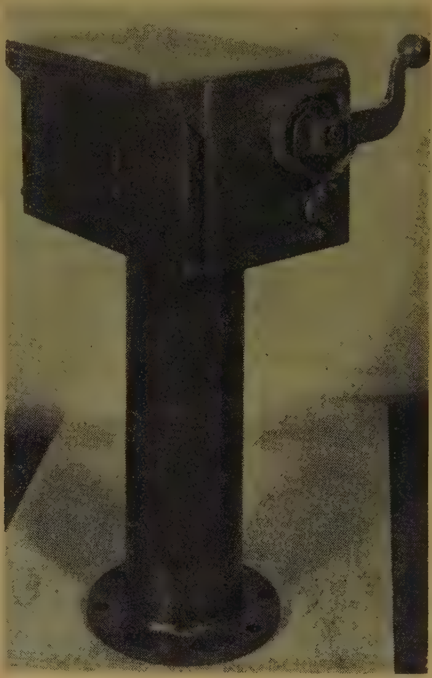


Figure 21. Waterproof master switch with spring-returned handle and seating facility for operator

Limitation of space does not permit discussion of special features required for different applications. Whatever control functions, however, are provided, emphasis should be placed on simplicity of design to obtain the desired results with a minimum amount of apparatus.

A New Approach to Probability Problems in Electrical Engineering

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THE PROBABILITY of simultaneous occurrences of events for which the individual rate of occurrence is known can be determined by the well-known binomial law. This law states that if n events have each a probability or rate of occurrence p, then the probabilities that 0, 1, 2, 3, etc., of the n events occur simultaneously, are given by the consecutive terms of the binomial function $[p+(1-p)]^n$. Therefore, the probability that r out of the n events occur simultaneously is

$$P_r = \frac{n!}{r!(n-r)!} p^r (1-p)^{n-r} \tag{1}$$

Applied to the probable frequency of simultaneous occurrences, this expression merely says that if each of the n events occupies a fraction p of the period of time under consideration, say a fraction p of a year, then the simultaneous occurrences of r events occupy a fraction P_r of the period under consideration. However, if P_r were found to be 1/52 according to the binomial law, then it would be known only that in a year the simultaneous occurrence of r out of the n events would occupy in the average a period of one week. It is not known, however, whether this simultaneous occurrence does happen once a year with a duration of one week, or once every two years with a duration of two weeks, or seven times a year with an average duration each time of one day.

It is obvious that knowledge of the duration and intervals between overlapping events can be of considerable importance in many practical cases. In problems of congestion, such as in telephone operation, not only the probability of the use of r out of n available channels is of interest, but also the average duration and frequency of such an occurrence. Similar problems arise in connection with intermittent use of electric power, such as by welding apparatus and elevators. In statistical problems of unavailability of equipment, average length and average intervals between occurrences of unavailability may be important, such as in cases of outages of lines, transformers, or generators of electric utilities.

The equations presented in the following, which apply to a variety of problems are based on an extensive analysis and on enumerations of a great variety of cases. The equations were found to be correct also for limiting or borderline cases. The methods of analysis and derivations of the equations are not presented, as they would require too much space and exceed the scope of this paper. For the same reason, discussion of some of the purely mathematical implications and extensions of these equations are omitted.

Special Case

Starting with the most simple case, assume that each of n events has a rate of occurrence $p = t/T$ so that each event occurs once in a period of T units of time and lasts each time t consecutive units of time. Further, be it assumed that the starting points of the events can change only in steps of multiples of the unit of time selected. Then the average duration, or overlap, of r simultaneous events out of n, expressed in the selected units of time is

$$t_r = \frac{T\frac{n!}{r!(n-r)!}t^r(T-t)^{n-r}}{\left\{T\frac{n!}{r!(n-r)!}t^r(T-t)^{n-r} - T\frac{n!}{r!(n-r)!}(t-1)^r[(T-t)-1]^{n-r}\right\}} \tag{2}$$

or simplified,

$$t_r = \frac{t^r(T-t)^{n-r}}{t^r(T-t)^{n-r}-(t-1)^r[(T-t)-1]^{n-r}} \tag{3}$$

From a standpoint of statistical mathematics, the numerator of equation 2 is the sum of all possible durations of the r–fold simultaneous events and the denominator is the sum of all repetitions of such r–fold simultaneous occurrences.

In terms analogous to the binomial law, the average overlaps may be expressed by the expansion of the function

$$\frac{[tx+(T-t)]^n}{[tx+(T-t)]^n-[(t-1)x+(T-t-1)]^n} \tag{4}$$

(Here x is merely a "generating function" for convenient identification of terms corresponding to exponent r on x. Only the coefficients are of interest.) The overlap t_r is then the ratio of the coefficients of x^r which occur in numerator and denominator. This ratio is identical with equations 2 and 3.

As stated before, equations 2 and 3 assume that the start of the events can shift only in multiples of the unit of time chosen. Of course, these steps can be made as small as desired by choosing a small enough time unit. The restrictions of unit steps can be completely removed if we substitute pT for t in equation 3 and pass to the limit. We then obtain for the overlap duration, t_r expressed as a fraction of the entire period T under consideration.

$$\lim_{T\to\infty} \frac{t_r}{T} = \frac{(1-p)p}{r+p(n-2r)} \tag{5}$$

With P_r from equation 1 and t_r from equation 3 or 5, we get the average interval T_r between r–fold simultaneous events by

$$T_r = t_r/P_r \tag{6}$$

The frequency of such overlaps is

$$f_r = \frac{1}{T_r} = P_r/t_r \tag{7}$$

Following is a practical application of these equations. A single-phase 60-cycle circuit is supplying 30 welders. On the average, each welder gives 25 random current impulses per minute to the circuit. Each impulse lasts on the average 5 cycles. What are the average durations and frequencies of the various possible amplitudes of impulses? Here t = 5 cycles, T = 1/25 minute or 144 cycles, and p = 5/144. The durations and frequencies of the impulses of various amplitudes, as calculated by equations 5 and 7, are given in Table I.

Values of currents up to six times the rating of one welder are quite frequent, indicating the possibility of serious voltage flicker. From the data in Table I, it is simple to calculate the equivalent current in the cable which produces the same heating of the conductor as the various current impulses.

More General Case

In the case discussed, all individual events had the same rate and duration of occurrence. Now consider the case where each of the n events occurs again once in

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Table I

r, Number of Welders Overlapping in Operation	t_r, Average Duration of Overlap in Cycles	f_r, Frequency of Overlap Per Minute
0	4.63	269
1	2.45	548
2	1.664	423
3	1.259	187.3
4	1.010	56.5
5	0.847	12.63
6	0.728	2.2
7	0.638	0.309
8	0.568	0.036
9	0.512	0.0035

the period T, but the duration of each event is different. The duration t has different lengths $a, b, c. . . .$ for the different events and the probabilities $p = t/T$ have consequently different values. Then, assuming first that each event can shift only in steps of multiples of the time unit, the average duration or overlaps of r-fold simultaneous occurrences, expressed in the units of time chosen, is given by the ratios of the coefficients in the numerator and denominator of the expanded function 9, that is,

$$t_r = \frac{gx^r}{hx^r} \tag{8}$$

where g and h are the coefficients of x^r in the numerator and denominator, respectively, of the expanded function

$$\frac{\Pi[tx+(T-t)]}{\Pi[tx+(T-t)]-\Pi[(t-1)x+(T-t-1)]} \tag{9}$$

Here Π signifies the products of the terms in the parentheses, with t assuming all the different values $a, b, c. . . .$ Written differently and assuming only three events with durations a, b, and c, the average duration of the various overlaps are given by the ratios of the coefficients of x^r in the numerator and denominator of the function

$$\frac{[ax+(T-a)][bx+(T-b)][cx+(T-c)]}{\{[ax+(T-a)][bx+(T-b)][cx+(T-c)]-[(a-1)x+(T-a-1)][(b-1)x+(T-b-1)][(c-1)x+(T-c-1)]\}} \tag{10}$$

Again the steps in which the events shift can be made as small as desired by choosing the unit of time suitably small. The limitation of shift in multiples of the time units can be eliminated completely by setting

$$p_a = \frac{a}{T},\ p_b = \frac{b}{T},\ p_c = \frac{c}{T}$$

$$q_a = 1-p_a,\ q_b = 1-p_b,\ q_c = 1-p_c$$

and passing to the limit $T \to \infty$. Then the overlaps of r-fold simultaneous occurrences expressed as a fraction of the period T are given by the ratios of the coefficients in the numerator and denominator of the terms x^r of the following function

$$\frac{(p_ax+q_a)(p_bx+q_b)(p_cx+q_c)}{(x+1)[(p_ax+q_a)(p_bx+q_b)+(p_ax+q_a)(p_cx+q_c)+(p_bx+q_b)(p_cx+q_c)]} \tag{11}$$

which can, of course, be generalized to any number of terms $n > 3$.

As an example, assume that a distributing station of an electric power company is supplied by three lines from different sources. According to statistical data each of the three lines can be expected to be forced out of service once a year. Due to the different nature of the lines, 48 hours are required to restore the first line to service, 24 hours are required for the second line, and 12 hours are required for the third line. Then, $a=48$, $b=24$, $c=12$, $T=8{,}760$, $p_a=48/8{,}760$, $p_b=24/8{,}760$, and $p_c=12/8{,}760$. The average duration and intervals of outages as calculated by equation 10 or 11 is given in Table II.

Events of Different Amplitudes

The preceding equations gave information on only the number of lines out of service. If the loads which can be carried by the three lines are different, then it may be more important to obtain information on the kilovolt-amperes out of service rather than the number of lines out of service. If the three lines of the preceding example have different load-carrying capacities, L_a, L_b, and L_c, then the average duration and intervals of loss of various amounts of kilovolt-amperes can be obtained from equation 10 by substituting x_a, x_b, and x_c for x. Then the average duration of the loss of L_a kilovolt-amperes is the ratio of the coefficients for x_a in the expanded numerators and denominators, the loss of L_b+L_c kilovolt-amperes is given by the coefficients of x_bx_c, the loss of $L_a+L_b+L_c$ kilovolt-amperes is given by the coefficients of $x_ax_bx_c$, and so on.

Completely General Case

Assume that an event n occurs f_n times during the period T where f_n may

Table II

Number of Lines out of Service	Average Duration of Outages in Hours	Average Interval of Outages in Years
0	2920	0.344
1	27.9	0.334
2	12.2	53.14
3	6.8	37,747

Table III

Number of Channels in Use	Average Length of Overlap in Seconds	Frequency of Overlap Per Second
3	1.02	0.206
2	0.88	0.501
1	0.74	0.392
0	0.62	0.097

be different for each event. The lengths of the f_n occurrences of the event n may be all different and so may be the intervals between occurrences. The individual lengths of these occurrences may be $t_{n\alpha}$ $(\alpha=1, 2, \ldots, f_n)$ units. The individual lengths of intervals between occurrences may be $\delta_{n\alpha}(\alpha=1, \ldots, f_n)$ units.

Let

$$\sum_{\alpha=1}^{\alpha=f_n} t_{n\alpha} = \theta_n$$

and

$$\sum_{\alpha=1}^{\alpha=f_n} \delta_{n\alpha} = \Delta_n$$

In other words θ_n is the sum of the lengths of the f_n different occurrences for the event n, and Δ_n is the sum of the intervals between occurrences. Therefore,

$$\theta_n + \Delta_n = T$$

Then, assuming again that the events can shift only in multiples of the unit of time, the average duration or overlap of $0, 1, 2, . . .n$, simultaneous occurrences is given by the ratios of the coefficients of the various terms x^r in the expansion of the function.

$$\frac{\prod_{n=1}^{n}[\theta_nx+\Delta_n]}{\prod_{n=1}^{n}[\theta_nx+\Delta_n]-\prod_{n=1}^{n}[(\theta_n-f_n)x+(\Delta_n-f_n)]} \tag{12}$$

It is of special interest that in the numerator as in the denominator, the individual lengths of occurrences for each event and the order of occurrences have no effect on the result, but only the sums of the lengths of occurrences and of the intervals and the frequency of occurrence, or number of repetitions are of importance. This fact, of course, simplifies the application of this general formula considerably. Finally, we can remove the restriction of shift of events in multiples of the time unit by substituting p_nT for θ_n and q_nT for Δ_n and, furthermore, setting $p_n/f_n = v_n$ and $q_n/f_n = w_n$.

When we pass to the limit $T \to \infty$ we obtain for the duration of the average

overlaps expressed as fractions of the period T,

$$\lim_{T \to \infty} \frac{t_r}{T} = \frac{\left(\overset{r}{\Pi} v_n \times \overset{n-r}{\Pi} w_n\right)}{\Sigma\left(\overset{r-1}{\Pi} v_n \times \overset{n-r}{\Pi} w_n\right) + \Sigma\left(\overset{r}{\Pi} v_n \times \overset{n-r-1}{\Pi} w_n\right)} \qquad (13)$$

where the superscripts on the product symbols designate the number of terms in the product. Each double product contains each index of one to n not more than once. The summations are taken over all different combinations of n subscripts or indices in two groups of r and $(n-r)$, each in the numerator, and of $(r-1)$, and $(n-r)$, and of r, and $(n-r-1)$ each in the denominator corresponding to the double products.

If all values of v_n are identical and, consequently, all values of w_n are identical, equation 13 reduces to

$$\lim_{T \to \infty} \frac{t_r}{T} = \frac{vw}{mw+(n-m)v} \qquad (14)$$

If $f=1$, then $v=p$ and $w=q=1-p$, and equation 14 reduces to equation 5.

As an example of the general case, assume that of three available channels the first is in use once in the period T of 10 seconds and the duration of use is 5 seconds. The second channel is in use twice in each period T of 10 seconds and the two durations of use, 2 and 5 seconds, occur with equal frequency. The third channel is in use three times in each period T of 10 seconds and the equally likely durations of these three uses are 1, 2, and 3 seconds, respectively. Then the average durations and frequencies of overlaps, as calculated by equation 12, are given in Table III.

This example, admittedly artificial to illustrate the general formula, is typical of problems which can occur in telephone practice and in other fields of electrical engineering. Still further generalization is possible, and has been carried out though not elaborated here, along two general lines. First, it is possible to work out a method—but not a formula—to express the duration and interval between recurrences of "r or more" simultaneous occurrences and to further extend it to "w or more" kilovolt-amperes, say, when individual events are of unequal capacity. Secondly, it is possible to determine the deviations from the mean of the durations and intervals given by the above formulas.

The methods and equations as outlined have helped considerably in solving practical problems which could not be solved by equations previously available.

Excitation, Control, and Cooling of Ignitron Tubes

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APPLICATION DATA for mercury pool tubes or tanks must include information covering rectifier capacity, control characteristics, excitation requirements, losses, cooling requirements, and other essential characteristics.

The factors affecting rectifier capacity, the relations between these factors, and the procedure for specifying rectifier capacity have been set forth in a companion paper. This paper presents a description of the excitation, control, and cooling characteristics of typical mercury-arc tubes of the ignitron type and outlines a procedure for specifying these characteristics, together with the limitations and operating requirements which must be considered.

Excitation

The excitation system must establish and maintain the conditions required for arc conduction. To do this it must perform three functions:

1. Start a cathode spot on the cathode pool.
2. Maintain the cathode spot.
3. Extend the ionization to the anode region.

Ignitors

In rectifiers of the ignitron type, a cathode spot is started each cycle by the ignitor. In order to perform the starting function successfully, the ignitor must be supplied with sufficient energy to assure reliable operation. The instantaneous volt-ampere requirements of the ignitor depend upon its construction, composition, and treatment. Individual ignitors vary in resistance and starting characteristics because of small differences introduced during manufacture or operation. In order to assure reliable ignitor operation, definite specifications have been set up for ignitor performance. Acceptable ignitors must fire at peak voltages and currents within the specified maximum values.

The ignitor excitation requirements are best described by a volt-ampere characteristic of the type shown in Figure 1. The "minimum" curve gives the instantaneous volt-amperes that this excitation circuit should deliver. It must be understood that an ample margin should be allowed between the "minimum" curve and the maximum volt-amperes required to fire acceptable ignitors. The "maximum" curve indicates the highest value of instantaneous volt-amperes that the firing circuit may deliver without damage to the ignitor or seal assembly.

A wide variation in the resistance of individual ignitors is permissible, and the dotted curves show typical characteristics for ignitors having resistances ranging from 2 to 100 ohms. The peak short-circuit current and open-circuit voltage indicated by the solid curves are particularly useful in specifying the excitation circuit requirements. Further limitations and requirements of the ignitors are indicated in Table I.

Auxiliary Anodes

In order to minimize the power consumed by the ignitors and also the physical size of the firing circuit components the ignitor volt-amperes are usually sustained for only a few electrical degrees (from 10 to 25 degrees in most circuits). Auxiliary excitation or relieving anodes, therefore, are used to perform the following functions:

1. Maintain the cathode spot for the required period. Many circuits require that excitation be maintained for 90 degrees in order to assure reliable operation at light load.
2. Furnish additional energy for ionization of the arc space.
3. Reduce the duty on rectifying elements in the firing circuit by providing a shunt path.

Table I. Ignitor Limitations and Requirements

Maximum rms current allowed, amperes	15
Maximum avg current allowed, amperes	3.0
Maximum instantaneous inverse voltage allowed, volts	5

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These auxiliary anodes may be located either in the open arc path or in a shielded space adjacent to the cathode pool. They are usually excited from a sinusoidal source. Typical wave forms are shown in Figure 2.

The inverse voltage applied to the excitation anode must be held to a low value in order to limit the positive ion bombardment and avoid arc-backs. The maximum allowable inverse voltage depends upon the amount of shielding around the excitation anode and the time of application of the inverse voltage, that is, whether the inverse voltage is applied during the main anode conduction or during the main anode inverse period. By proper phasing of the excitation anode voltage with respect to the ignitor firing point, the excitation anode conduction period may often be fixed so that inverse voltage is applied to the excitation anode only during the main anode inverse period. Where this is not possible, some means, such as a rectifying element), must be provided to limit the inverse voltage which may be applied. Further limitations and requirements are indicated in Table II.

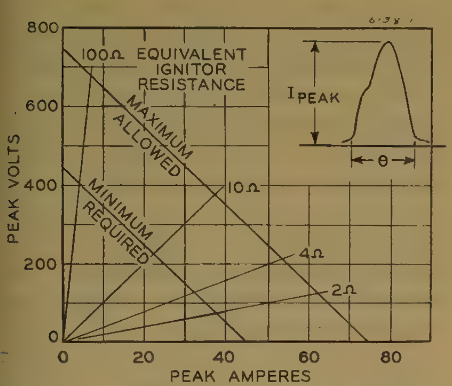

Figure 1. Ignitor instantaneous volt-ampere requirements

Pulse duration $\theta = 10$ to 15 degrees at 25 to 60 cycles

GRIDS

While the grids are also a part of the excitation system, in that they help to establish the conditions for conduction, this is only incidental to their control action. Their characteristics are covered in the section on control.

Control

In an ignitron, control of conduction may be obtained by means of either ignitors or grids. The ignitor provides the only means of control in many small tubes which are built without grids. In such tubes, the ignitor usually provides effective control over the full range of operation. However, most larger tubes are equipped with one or more grids.

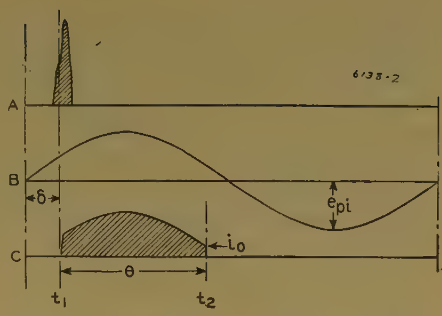

Figure 2. Excitation anode wave forms

A—Ignitor current
B—Excitation anode voltage
C—Excitation anode current

The primary function of the grid is to control both the arc phenomena in the neighborhood of the anode and the starting of anode conduction. While the establishment of suitable arc conditions at the anode is essential to the action of the rectifier and affects the rectifier capacity directly, the control function of the grid is the more important from an application standpoint.

The grid action depends directly upon the electronic processes taking place in the rectifier throughout the cycle. These are well illustrated by the oscillographic observations shown in Figure 3. These observations were made on a pumped ignitron operating in a six-tank unit at $E_{do} =$ 1,294 volts and I_d = 2,000 amperes load current. In order to facilitate analysis of the action, the grids were excited from a rectangular voltage wave of 125 volts peak through 500 ohm resistors.

Referring to Figure 3, main anode conduction starts at time t_0 with a positive grid current. At time t_1, before the end of conduction, the voltage applied to the grid circuit is made negative. The grid potential remains substantially unchanged although the grid current is reversed. The positive ion current collected by the grid is limited only by the resistance of the grid circuit. Main anode conduction ends at time t_2. However, the grid current is sustained as the grid continues to collect positive ions from the residual ionization. As deionization continues, the intensity of ionization finally, at time t_3, falls below that required to sustain the full positive ion grid current, and the grid current then falls off along a

Table II. Excitation Anode Limitations and Requirements

Maximum rms current allowed, amperes	10
Maximum avg., current allowed amperes	5
Minimum instantaneous current required to maintain cathode spot, amperes	3
Maximum peak inverse voltage allowed during conduction, volts	25
Maximum peak inverse voltage allowed during inverse, volts	150

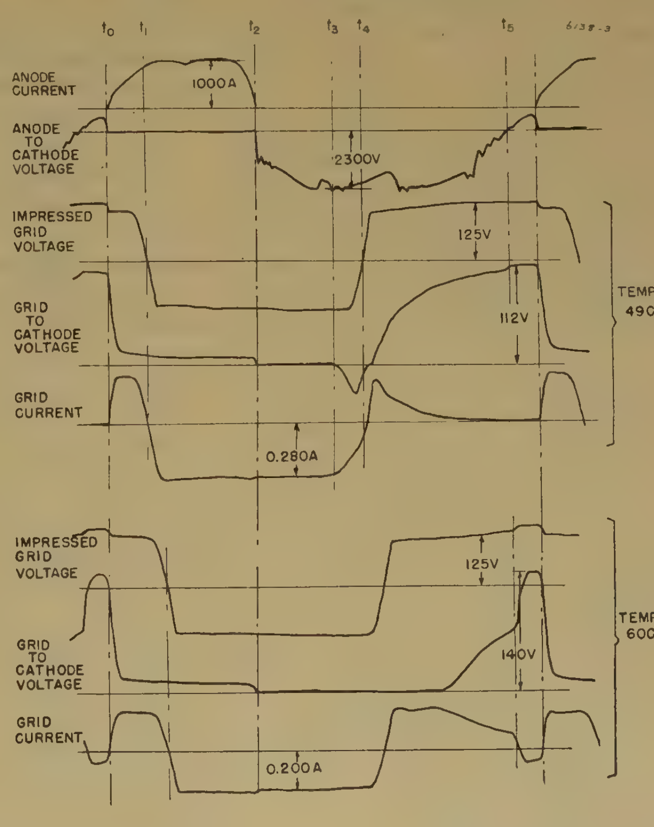

Figure 3. Grid characteristics on pumped ignitron with single grid

decrement curve as deionization progresses. At time t_4, the voltage applied to the grid circuit again becomes positive. The grid now collects electrons and since these are more mobile than the positive ions and can be drawn from the entire arc space, the grid current again reaches the limiting value fixed by resistance. However, the deionization process continues and this current also falls off on a decrement.

Late in the unverse cycle, the main anode becomes positive at time t_5. Since the rectifier is operating with phase control and the ignitor has not yet fired, conduction does not begin. However, the grid again collects ions as it is made negative with respect to the anode. This action is quite pronounced when operating at 60 degrees centigrade temperature.

Many conclusions regarding the action may be deduced from these observations. Some of these are as follows:

1. Ionization persists throughout the inverse cycle.

2. The deionization time is a function of the control temperature.

3. The action of the grid assists the decay of ionization.

4. The grid shields the anode from the ionization which persists in the arc path during the inverse cycle.

Grid Characteristics

Most ignitron tubes with a single grid have a negative grid characteristic, that is, require a negative grid potential to prevent conduction. This negative grid characteristic is obtained because the arc path is usually quite open and a high level of ionization is provided at the grid by the ignitor and auxiliary anode arc.

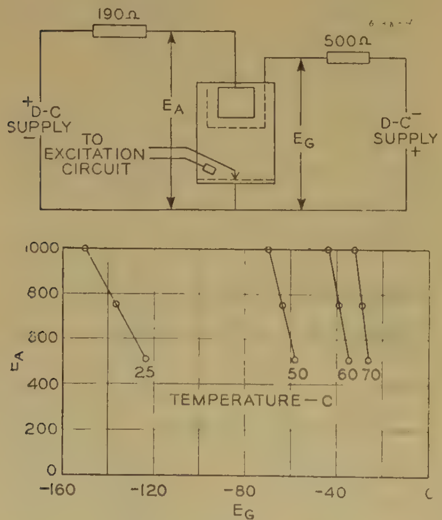

Figure 4. Grid blocking characteristic of pumped ignitron with single grid

E_g—Grid to cathode voltage at start of discharge

E_a—Anode to cathode voltage

Typical grid-blocking characteristics for a pumped ignitron having a single grid are shown in Figure 4. These characteristics apply where the ignitor is energized and the blocking action of the grid is utilized to prevent conduction while the rectifier is being placed in operation. The grid current which flows as a result of the ionization produced by the ignitor and excitation anode is shown in Figure 5.

Table III. Grid Limitations and Requirements

Minimum grid resistance allowed, ohms	500
Maximum grid resistance allowed, ohms	5,000
Maximum inverse voltage allowed during conduction, volts	150
Maximum inverse voltage allowed during inverse, volts	300
Minimum positive voltage required to fire, volts	50
Maximum positive voltage allowed, volts	300

The grid characteristics, when operating with phase control, are similar to those for blocking, except that the presence of the ionization from the conduction of the preceding cycle increases the negative voltage required on the grid to prevent arc-through, or loss of control. The slow deionization of a single grid tube is shown clearly in Figure 3. For this reason, a single grid does not provide effective control over a wide range of temperature and ionization. A wider range of control may often be obtained by the ignitor than by the grid in a single grid tube.

Table III gives a further list of typical grid limitations and requirements.

Where a wide range of phase control is required to obtain blocking action from no load to fault currents, full phase control at all loads, and accurate timing of firing, two or more grids are usually provided in the ignitron tube. With two grids, the outer grid shields the inner from the ionization in the arc path and permits more rapid and complete deionization of the space around the anode and the inner grid so that better control characteristics are obtained. A wide variety of firing characteristics may be obtained by such arrangements. These are generally similar to those for the single grid. However, further studies are necessary for a full treatment of their characteristics, limitations, and requirements.

Cooling and Temperature Control

Mercury-pool tubes must be operated within a specified temperature range in order to deliver rated power output, assure reliable operation, and prevent damage during faults. The temperature also affects the rectifier efficiency since the arc losses are a function of temperature.

The temperature regulating equipment which is provided for the purpose of establishing the desired temperature conditions must perform the following major functions:

1. Remove the heat produced by the arc losses.

2. Maintain the desired control temperature.

3. Maintain the desired temperature gradient on the anode seals.

4. Minimize corrosion.

Arc Losses

Most ignitron tubes have an arc drop ranging from 15 to 25 volts. The losses are approximately proportional to the load current although the arc drop increases somewhat with the load. The maximum cooling requirements are usually determined by the arc loss obtained when operating at the current required for the two-hour overload condition. Since the heat storage capacity of the usual tube structure is equivalent approximately to a thermal time constant of 15 to 30 minutes, overloads of short duration,

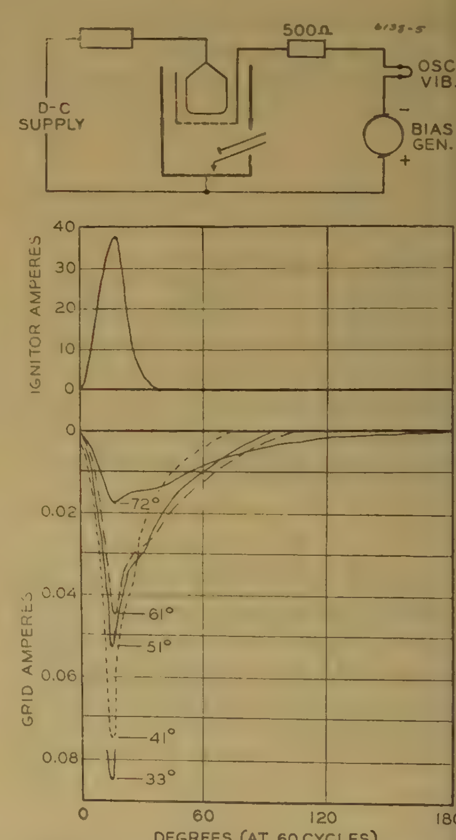

Figure 5. Grid current during blocking tests

Anode to cathode voltage = 0
Grid to cathode voltage = −300 volts
Temperature as shown

such as one minute, need not be considered in the determination of the cooling requirements.

Figure 6 shows a typical arc drop characteristic for a pumped ignitron operating in a unit of six tanks with a double-Y connection. This represents the usual mode of operation where $P = 3$, the conduction period is $(120° + u)$, and the peak tube current is three times the average tube current. An alternate connection which is sometimes used is the fork where $P = 6$, the conducting period is $(60° + u)$, and the peak tube current is six times the average tube current. The arc drop in this case will be higher because of the increased peak current and the losses, therefore, will be greater. Tests on two units show that the losses at the same load are increased approximately eight per cent over those obtained when operating double-Y. A similar correction factor, but having a different value, may be determined for other modes of operation.

Cooling

The majority of power rectifiers are cooled by water which is passed through suitable water jackets. Two general systems of water cooling may be employed. One is the direct raw water-cooling system in which water is taken from a suitable supply, passed directly over the cooling surfaces of the rectifier, and discharged. The other is the heat-exchanger cooling system in which a coolant is passed over the cooling surfaces of the rectifier, then cooled in a heat exchanger and recirculated.

Direct raw water cooling requires the least amount of equipment and, therefore, frequently is used on small units ranging from 75 kw to 300 kw. In such systems, tubes must operate over a wide temperature range with varying water flow. In order to conserve water, the tube water jackets usually are connected in series.

Heat-exchanger cooling systems are used for all the larger rectifiers where it is desired to operate the tubes at the optimum temperature and utilize their full capabilities. Pertinent information necessary in applying a heat exchanger to a rectifier unit is as follows:

1. Type and design of rectifier.
2. Water flow in gallons per minute.
3. Pressure drop through water jacket, in pounds.
4. Approximate arc losses in kilowatts.

Anode Heaters

It is necessary to maintain a temperature gradient at all times from the anode and grid seals to the tank, so as to prevent the condensation of mercury on the seal insulators. On equipment employing sealed tubes, with direct raw water cooling, usual service conditions are generally such that no heaters are required. Where sealed ignitrons operate under unusual service conditions which are unfavorable, anode heaters must be provided. Anode heaters generally are required on all pumped ignitrons. These heaters range up to 350 watts, depending upon the size of the tube and the service conditions.

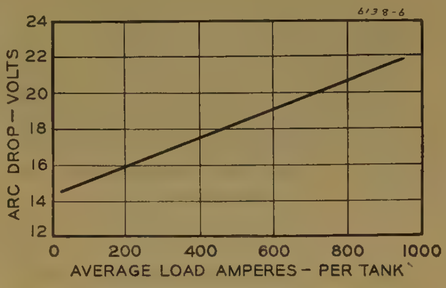

Figure 6. Typical arc-drop characteristic for pumped ignitron

Corrosion

The prevention of corrosion in rectifier cooling systems, involves two considerations, one, a proper choice of metals which come in contact with the water, and second, the control of the chemical constituents of the coolant. The sealed tubes employ stainless steel for the vacuum tank in order to insure vacuum tightness. A complete corrosion-resistant water chamber is readily obtained by providing water jackets of the same material. The stainless steel construction of sealed tubes makes them suitable for use in direct raw water cooling systems. Pumped rectifiers employ vacuum tanks made of boiler plate steel. They must be proved with copper cooling coils for use in direct raw water-cooling systems.

For heat exchanger systems, where a recirculating coolant is used, the materials in the associated piping which is connected to the rectifier should consist of malleable iron pipe and fittings. It is good practice that the recirculating water be treated with sodium chromate (Na_2-CrO_4), using either distilled or an approved raw water.

Conclusions

The factors which must be considered in the application of mercury-pool tubes and the design of rectifier units have been described. Also a procedure for specifying their characteristics, limitations, and requirements has been outlined.

A full knowledge of the characteristics, limitations, and requirements of the ignitron tube is essential to the design of a successful and economical rectifier unit inasmuch as the auxiliaries required for excitation and temperature control are an important part of the unit.

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Dimensionless Analysis of Servomechanisms by Electrical Analogy

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THE MECHANICAL transients analyzer[1,2] provides a fast and simple method of obtaining the complete solutions to the performance equations of servomechanisms under any type of operating condition. It has been found to be applicable to the detailed analysis of specific systems. Also, because of the great rapidity with which parameters of a basic system can be changed once that system is set up, the analyzer is admirably suited to generalized studies of various types of systems. This paper presents a general study that has been made of variable-voltage angular-position servomechanisms. The data give dominant system frequencies and damping rates for a two time delay system with simple error control and for a wide range of parameters. Similar data are presented for error control plus optimum anticipation time constants, together with other performance characteristics. Because of the many curves required for the same type of analysis of the several variations of resistance-capacitance-feedback and other forms of antihunt or damping devices, they will not be treated here.

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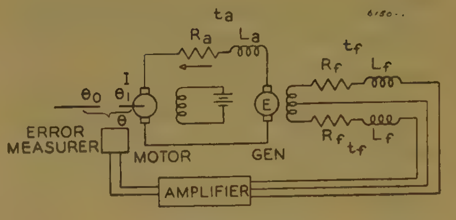


Figure 1. Schematic diagram for variable-voltage angular-position servomechanism (see appendix)

The range of parameters chosen is sufficiently wide to cover all practical types of variable-voltage servo applications. The performance of a given system or the choice of parameters for optimum performance can be quickly obtained from the dimensionless data on the curves.

Description of System Studied

A schematic diagram of the system studied, together with the definition of all parameters and the system equations, is given in Figure 1. In this system two time delays are included. One of these, t_a, can only represent the armature circuit time delay. However, if the field time delay of the system being studied is negligible, t_f can represent any other time delay back in the system. The appendix gives the dimensionless parameters upon which the generalized study is based.[3] These include the undamped natural frequency ω_n with no time delays, the ratio, r, of actual to critical damping with no time delays, and the time constants put in dimensionless form through multiplication by ω_n. In the accompanying curves, the performance solutions plotted as functions of the above parameters are actual dominant system frequencies, dominant system damping rates, and crest values of the error angle. The first two are applicable to any form of disturbance or motion to be followed. As shown in the appendix, system frequencies are put in dimensionless form by expressing them as a fraction of ω_n. Damping is given in per cent per cycle. Since a suddenly applied constant velocity is considered to be the most basic type of disturbance function for such systems, the actual solutions were obtained for this case and the crest error angles, θ_{max}, or transient overshoots, were put in dimensionless form by expressing them as a function of the steady-state error angle, θ_{ss}, for such disturbance (see appendix).

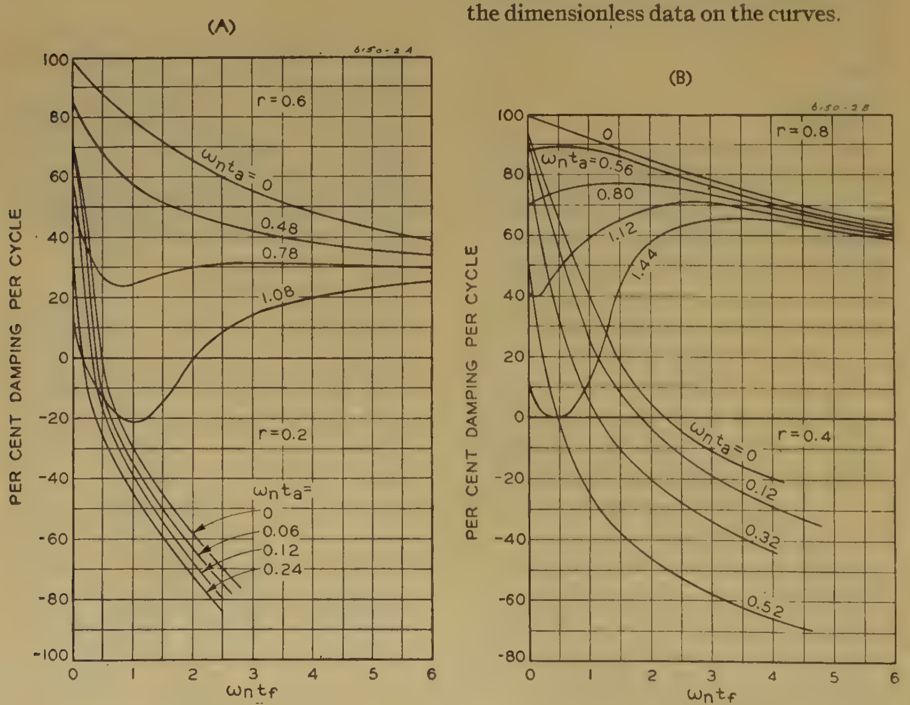
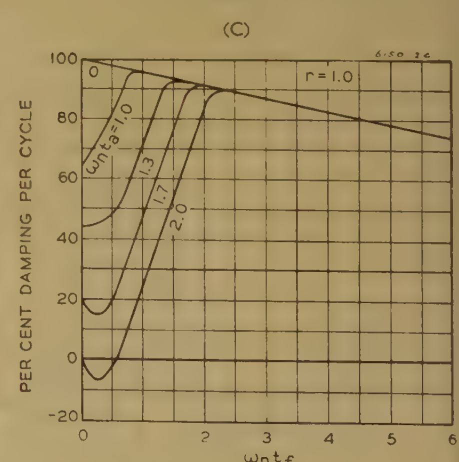
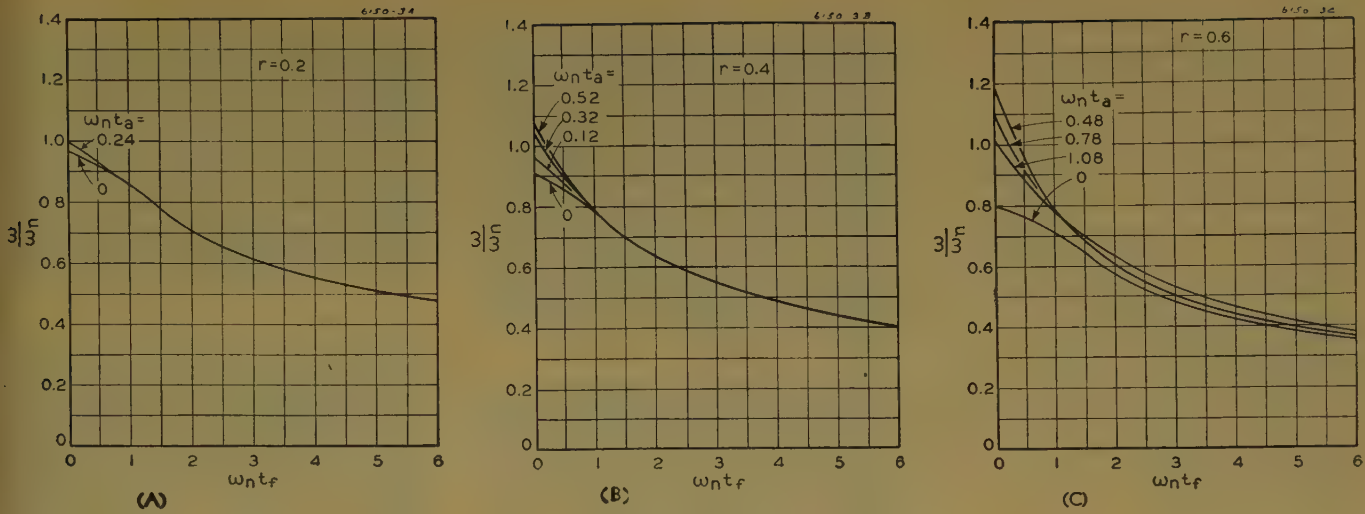


Figure 2. Dimensionless curves of per cent damping for variable-voltage (two time delays), angular-position servomechanisms with simple error control

A. $r=0.2$ and $r=0.6$
B. $r=0.4$ and $r=0.8$
C. $r=1.0$





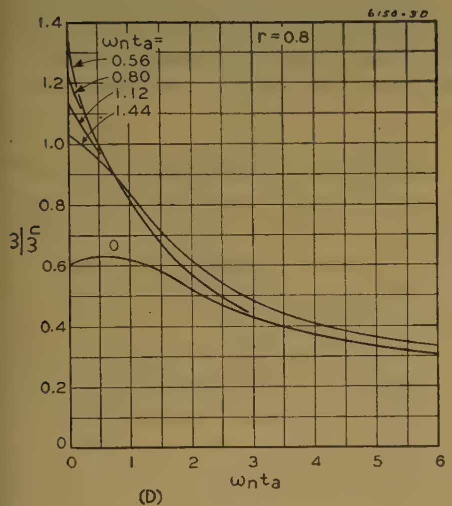
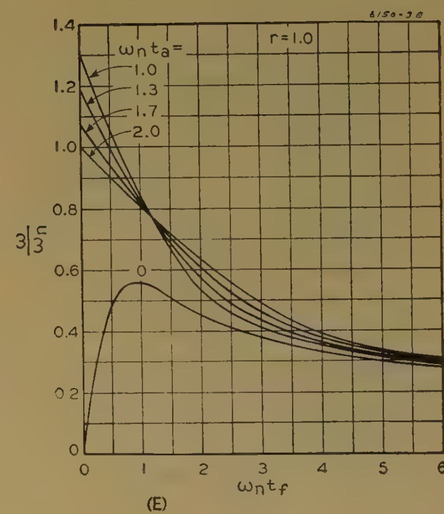


Figure 3. Dimensionless system frequencies for variable-voltage angular-position servomechanisms with simple error control

A. $r=0.2$
B. $r=0.4$
C. $r=0.6$
D. $r=0.8$
E. $r=1.0$



The method of studying such systems on the transients analyzer is discussed in detail in reference 1. The basic torque equation of Figure 1 is represented as an electric circuit with suitable circuit constants and amplifiers. The exact circuits and analogies used are given in the above reference. The complete transient solutions were observed or photographed on the screen of a cathode-ray oscilloscope. In reading and recording data either from the oscilloscope screen or the photographs, precautions were taken to obtain about five per cent accuracy. Numerous check calculations were made for the case of simple error control, all of which checked well within ten per cent. It should be pointed out that the data presented represent the equivalent of the solution of several thousand servo-equations. The greatest number of solutions would be required in obtaining the optimum anticipation which is quite simple with the analyzer.

Analysis of Data

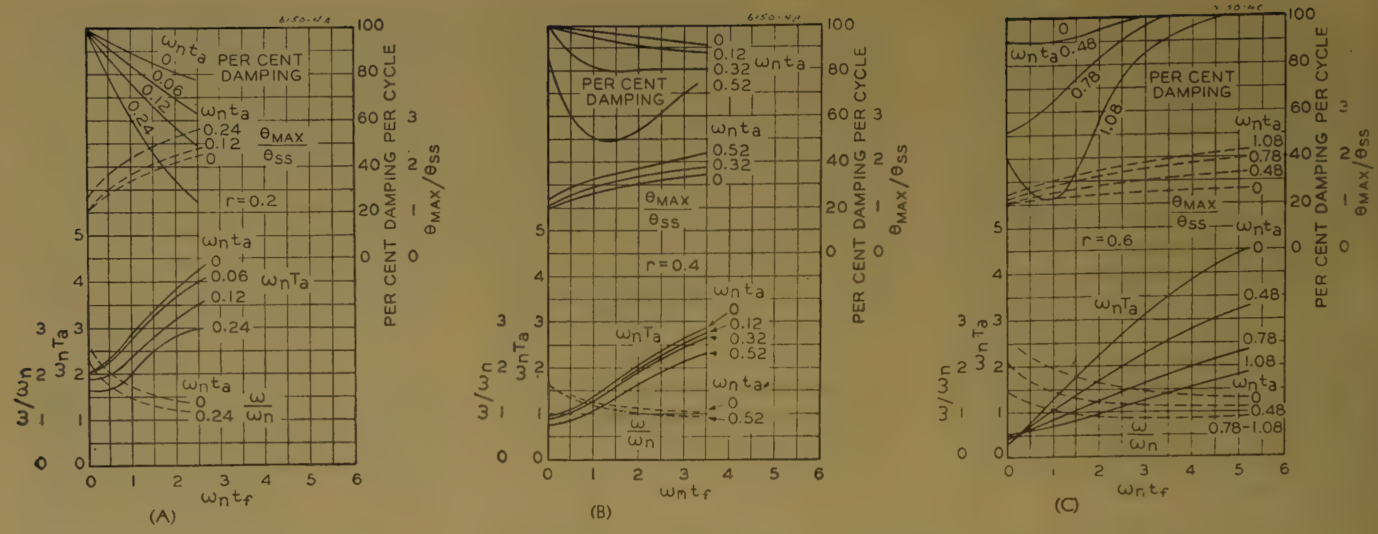
The information presented in Figures 2, 3, and 4 is in dimensionless form. Therefore, any physical system of the type shown in Figure 1 is represented by the curves of Figures 2, 3, and 4 regardless of the actual numerical values of its parameters.

Figure 2 shows curves of per cent damping per cycle plotted as a function of $\omega_f t_n$ for different values of $\omega_n t_a$, and r equal to 0.2, 0.4, 0.6, 0.8, and 1.0. All symbols used are as defined in the appendix. In Figure 3 the ratio of actual frequency, ω, to the undamped natural frequency, ω_n, is also plotted as a function of $\omega_n t_f$ for various values of $\omega_n t_a$, and r equal to 0.2, 0.4, 0.6, 0.8, and 1.0. The data of Figure 4 were obtained by choosing an anticipation time, $\omega_n T_a$, to provide the best compromise between damping, response, and overshoot. These data give per cent damping per cycle, the ratio of maximum angular error, θ_{max}, to the steady-state angular error for a constant input velocity, θ_{ss}, the ratio of actual to undamped frequency, ω/ω_n, and the optimum value of anticipation time, $\omega_n T_a$, as a function of the two system time delays. Separate curves are given for r equal to 0.2, 0.4, 0.6, 0.8, and 1.0. Figure 5 shows photographs of some of the typical solutions obtained in deriving the curves of Figure 4. The ordinate is angular error θ, and the abscissa, time.

A study of Figures 2, 3, and 4 shows that any general conclusions drawn may have exceptions at particular points. Thus, although frequency and damping generally decrease with increase of $\omega_n t_f$, in Figure 3C, for $\omega_n t_a = 0$, frequency increases with increase of $\omega_n t_f$ up to about 1.0, and in Figure 2C, damping increases for a number of values of $\omega_n t_a$ for increase in $\omega_n t_f$ to 1.0 or better.

In order to have satisfactory performance of the simple-error type of servomechanism, it can be seen in general in Figures 2 and 3 that large values of r and small values of $\omega_n t_a$ and $\omega_n t_f$ are required. The exception to the above statement is found at values of r of 0.8 and 1.0 where it is possible to increase the damping to a satisfactory level by increasing $\omega_n t_f$, although there is a sacrifice in frequency ratio.

The use of an error rate or anticipation term in the controller usually results in a well-damped system with good frequency response when r is small and the time delays $\omega_n t_a$ and $\omega_n t_f$ are large. This is well illustrated by Figure 4 and by actual photographs of such a case, Figures 5A and B. However, for the practical range of time delays for r equal to 0.2, damping decreases for all cases of increasing time delays. Difficulty is encountered in obtain-



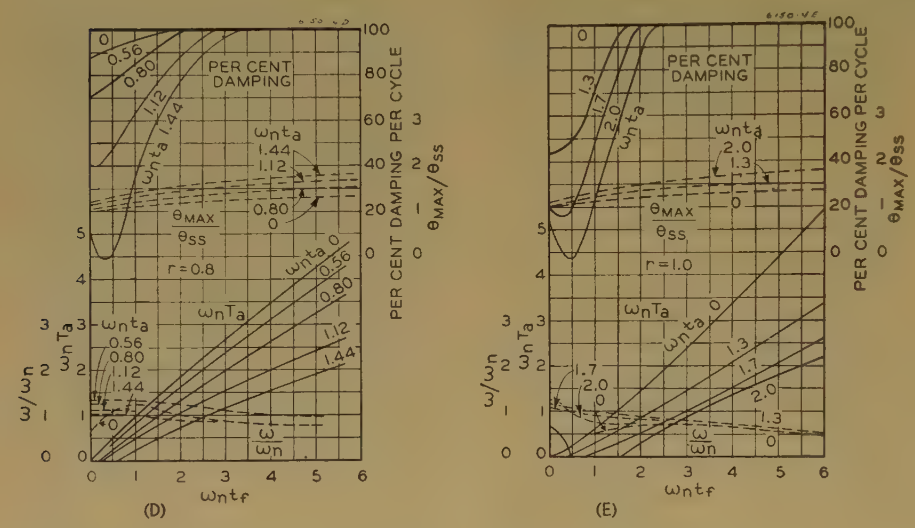


Figure 4. Dimensionless performance curves for variable-voltage angular-position servomechanisms with optimum anticipation time ($\omega_n T_a$)

A. $r=0.2$
B. $r=0.4$
C. $r=0.6$
D. $r=0.8$
E. $r=1.0$

ing sufficient damping at r equal to 0.6, 0.8, and 1.0 for large values of $\omega_n t_a$ and small values of $\omega_n t_f$. The photographic record of Figure 5C and D shows this. In Figure 5D the principal effect of anticipation is to reduce the magnitude of the oscillation with little or no increase in damping. It should be recognized immediately that this is not a common operating point, for the armature time delay is rarely even of the same order of magnitude as the field delay.

In general, the optimum anticipation time, $\omega_n T_a$, increases; the frequency ratio, ω/ω_n, decreases; and the peak ratio, θ_{max}/θ_{ss}, increases with increase in $\omega_n t_f$.

Remembering that high values of r bring correspondingly higher steady-state errors for a constant input velocity and greater power loss, the advantage of anticipatory control over the simple error control is apparent.

The curves of Figures 2, 3, and 4 have been smoothed so that some of the small kinks, caused by the large number of roots in the differential equations, have been omitted. At all points, however, the curves are well within an accuracy of ten per cent. There is no difficulty in choosing the dominant frequency in most cases. However, where a second frequency is pronounced, the ω chosen was that which corresponded to the time for the initial overshoot to first return to the steady-state angular error, θ_{ss}.

Typical Example

To clearly illustrate how easily the data in Figures 2, 3, and 4 can be used, a typical variable-voltage angular-position servomechanism problem is worked out below.

The parameter values are as follows:

$I=10^{-1}$ pound-inch-second^2
$K=5$ inch-pounds per radian
$R_a=6$ ohms
$L_a=0.10$ henry
$R_f=100$ ohms
$L_f=15$ henries
$b=4$ inch-pounds per ampere
$g=0.6$ volt per radian per second

$$\omega_n=\sqrt{\frac{K}{I}}=7.07 \text{ radians per second}$$

$$c=\frac{bg}{R_a}=0.4 \text{ inch-pound per radian per second}$$

$$r=\frac{c}{2\sqrt{KI}}=0.283$$

$\omega_n t_f=1.06$
$\omega_n t_a=0.12$

In this case all parameters as I, K, and b that are a function of shaft speed are given at the motor shaft.

Any shaft geared to the motor could easily have been used as a reference instead, as long as all the values were given at that shaft. At times this may be desirable.

Referring to Figure 4 and interpolating between Figure 4A and B for $r = 0.28$ gives the following data:

Optimum $\omega_n T_a=2.0$
Per cent damping per cycle$=83$
$\omega/\omega_n=1.4$
$\theta_{max}/\theta_{ss}=1.7$

Therefore, if in the actual system T_a is made $2.0/7.07=0.283$ second, the actual system will have a frequency of $1.4\times 7.07=10$ radians per second, and a maximum angular error at the motor shaft for a suddenly applied constant velocity, ω_1, of

$$1.7\times\frac{2r\omega_1}{\omega_n}=0.137\omega_1$$

The damping per cycle of the system is 83 per cent.

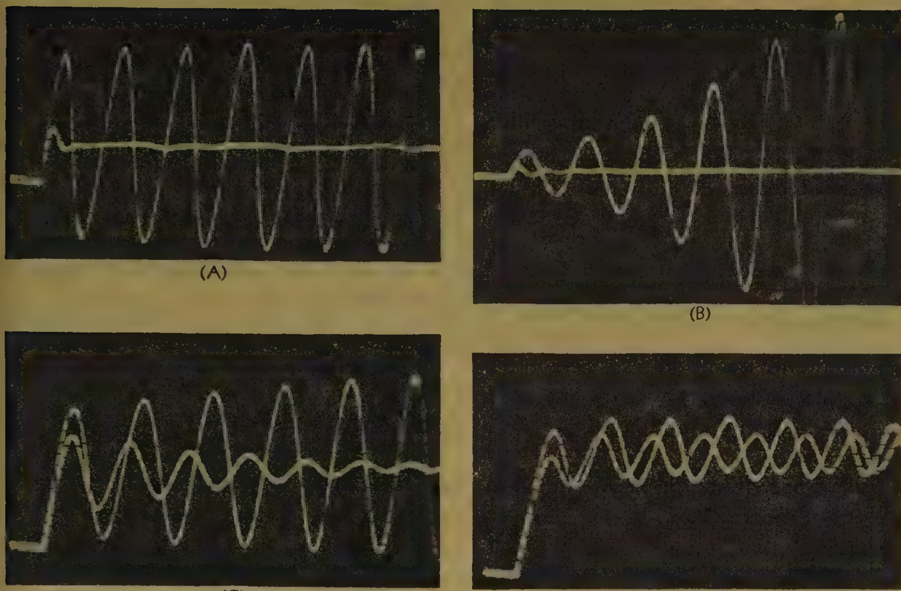


Figure 5. Typical solutions as obtained with the mechanical-transients analyzer

A. $r=0.2$, $\omega_n t_a=0$, $\omega_n t_f=0.53$, $\omega_n T_a=2.19$
B. $r=0.2$, $\omega_n t_a=0.24$, $\omega_n t_f=0.53$, $\omega_n T_a=1.64$
C. $r=0.6$, $\omega_n t_a=1.08$, $\omega_n t_f=1.06$, $\omega_n T_a=0.6$
D. $r=1.0$, $\omega_n t_a=2.0$, $\omega_n t_f=0.53$, $\omega_n T_a=0.55$

Conclusions

The data presented provides in clear form the solutions to any variable-voltage angular-position servomechanism with two time delays for either simple error control or error plus error rate (anticipatory) control. The solutions include system frequencies and per cent damping per cycle that are applicable to any system disturbance, and maximum overshoot data are provided for the case of suddenly applied constant velocity. A typical practical example of the use of the data is given.

It should be emphasized that the solutions are directly applicable only to systems that are approximately linear, although the correct order of magnitude of results will be obtained even for systems that materially depart from linearity.

Appendix

Equations for Figure 1

t_a = armature time constant
t_f = field time constant
T_a = anticipation time constant
K = stiffness constant

Equation for armature circuit:

$$E_g=R_a i+L_a p i+g p\theta_0;\quad p=\frac{d}{dt}$$

Equation for motor torque:

$$bi=Ip^2\theta_0$$

Giving the mechanical torque equation

$$M_c=\frac{bE_g}{R_a}=\frac{L_a I}{R_a}p^3\theta_0+Ip^2\theta_0+\frac{bg}{R_a}p\theta_0$$

$$M_c=(1+t_a p)Ip^2\theta_0+cp\theta_0$$

Controller torque M_c for simple error control:

$$M_c=\frac{K(\theta_1-\theta_0)}{(1+t_f p)}$$

Controller torque for error control plus anticipation:

$$M_c=\frac{K(1+T_a p)(\theta_1-\theta_0)}{(1+t_f p)}$$

Dimensionless parameters for generalized analysis:

$\omega_n=\sqrt{K/I}$ = undamped natural frequency with no time delays

$r=\frac{c}{2\sqrt{IK}}$ = ratio of actual to critical damping with no time delays

Dimensionless time constants $=\omega_n t$
Dimensionless system frequencies $=\omega/\omega_n$
ω_1 = suddenly applied constant velocity

$\theta_{ss}=\frac{2r\omega_1}{\omega_n}$ = steady-state error for suddenly applied constant velocity

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Circuit Cushioning of Gas-Filled Grid-Controlled Rectifiers

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Synopsis: In certain common rectifier and inverter circuits, gas-filled rectifier tubes supplying highly inductive loads with firing delayed by grid control, are subjected to the application of high rates of rise of initial inverse voltage. This phenomenon results in short tube life because of the sputtering of anode material by the impact of residual ions attracted at high velocity to the negatively charged anode with consequent gas cleanup. The paper describes a method of slowing down or cushioning this rate of rise of initial inverse voltage. A small resistance and capacitance circuit connected between cathode and anode of each tube delays the voltage rise the few microseconds necessary to eliminate gas cleanup. Life test data and practical applications are cited.

GAS CLEANUP, or gradual disappearance of the gas filling of gaseous discharge tubes was noted by Plucker in 1858, and later studied more extensively.[1] The present work was undertaken to explore the phenomenon as related to hot-cathode xenon- and argon-filled grid-controlled rectifiers.

The exploration was prompted by a reported tube life of 40,000 hours without

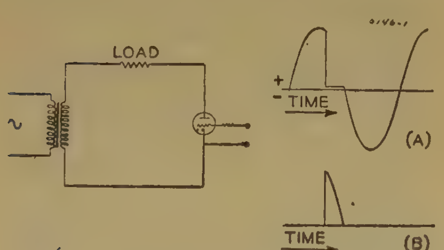

Figure 1. Circuit of test 1 with oscillograms of tube voltage and current

A—Voltage across tube
B—Current through tube

cleanup in one circuit, contrasted with a circuit which consistently caused cleanup within 250 hours.

Cleanup Test Data

Life tests were initiated in 1936 under the tube operating conditions listed in Table I, using a tube type which tests on resistive load under widely varied conditions had shown to be free from gas cleanup and dependable for lives in excess of 3,000 hours.

Before this work was undertaken, it was thought that cleanup might accompany rapid ionization.

Test 1 was designed to give rapid current buildup by reason of a low impedance power supply and transformer. Tube current rose abruptly from zero to 90 amperes each cycle. As the test was at overload, the life obtained was considered normal. Tests after failure showed that cleanup did not occur. Hence rapid buildup of ionization is not cause for cleanup at least up to rates of current rise of several million amperes per second.

Test 2 applied high initial inverse voltage caused by the current carry-over forced by the inductive load, shown in the diagrammatic oscillograms of tube voltage and current in Figure 2. Here no rapid buildup of ionization occurred, but in each cycle, at the end of conduction, a high inverse voltage abruptly appeared. This initial inverse voltage is indicated on the oscillogram of the tube voltage in Figure 2. With this circuit condition rapid cleanup occurred.

In test 3 the load inductance was made larger to prohibit appreciable change of current throughout the cycle. At the end of each cycle, the inductance generated the instantaneous voltage necessary to overcome negative supply voltage, thus forcing current to flow through the conducting tube until relieved by the firing of its successor. Current flowed through each tube in square blocks, thereby maintaining full current just prior to applying the initial inverse voltage. This test caused drastic cleanup.

Table I further indicates by test 4 that lowered initial inverse voltage alleviates cleanup somewhat, whereas increased frequency in test 5 aggravated cleanup.

As a result of this and other data, and numerous installations where it was shown by test that cleanup did not occur, it was concluded that gas cleanup would not occur, provided:

1. The circuit did not apply each cycle high initial inverse voltage immediately following substantial current conduction.

2. No sizable ion current was drawn to the grid during tube conduction nor was a sizable electron current drawn to the grid during periods of high inverse voltage on the anode.

3. There were no parasitic glow discharges or tube abnormalities.

Items 2 and 3 had been found some years previously to be possible causes for cleanup. In the tests reported in Table I, suitable grid resistors were used (approximately 50,000 ohms) to eliminate cause 2. Many previous tests of this tube type on resistance loads under various conditions had demonstrated that the tubes were free of cleanup from cause 3.

Since 1937, limitation 1, as part of the tube application notes of one company, prohibited many commercial applications of gas-filled tubes. Desiring to remove this restriction, the details of the abrupt transfer from conducting current to withstanding inverse voltage were probed further.

Commutation

This transition for the circuit shown in Figure 3 is depicted with expanded time scale in Figure 4. The process of current commutation commences at instant t_s as the tube about to enter construction 2 is fired by grid action. Its anode voltage falls precipitously and conduction current arises. However, leaving tube 1 continues conduction by reason of the inductance of the supply and transformer, shown schematically as equivalent leakage inductances L_1. Similar inductance L_2 momentarily limits the growth of current through entering tube 2. With both tubes conducting concurrently, both anodes are at the same potential and

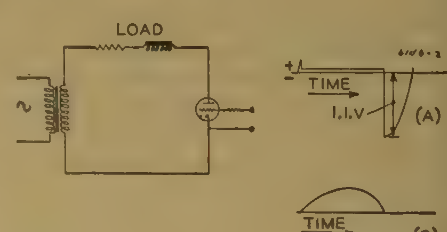

Figure 2. Circuit of test 2 with oscillograms of tube voltage and current

A—Voltage across tube
B—Current through tube

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the transformer is temporarily short-circuited. Full voltage during this period is applied to L_1 and L_2 in a direction to increase current in L_2 and depress it in L_1. By the time t_1, current through leaving tube 1 has been brought to zero and load current transferred to entering tube 2. Rectifier action prohibits reverse current, therefore the rate of change of current through L_1 disappears and the voltage across it collapses from maximum value to zero. This applies the full negative voltage of the winding to the leaving anode. Simultaneously, load current, which cannot change rapidly because of the high load inductance, becomes the only current flowing through L_2, causing voltage across L_2 to disappear and further increasing the initial inverse voltage on leaving tube 1. Stray capacitances and core losses in the various inductances prevent the initial inverse voltage from appearing absolutely instantly. It is not until t_2, that full initial inverse voltage appears. However, on circuits of a few kilowatt capacity, voltage often appears at a rate of over 100,000,000 volts per second. All of the commutation events just described occur in a time so small compared to 60-cycle waves that they often go unnoticed.

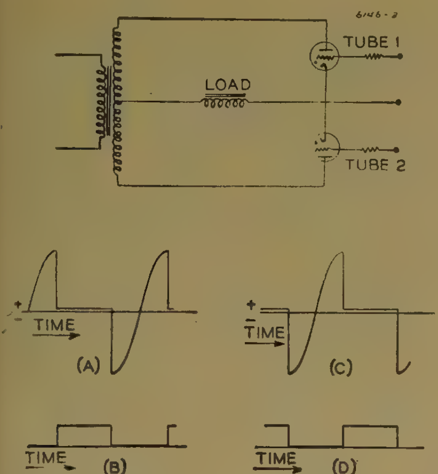

Figure 3. Circuit of tests 3, 4, 5, and 6 with oscillograms of tube voltages and currents

A—Voltage across tube 1
B—Current through tube 1
C—Voltage across tube 2
D—Current through tube 2

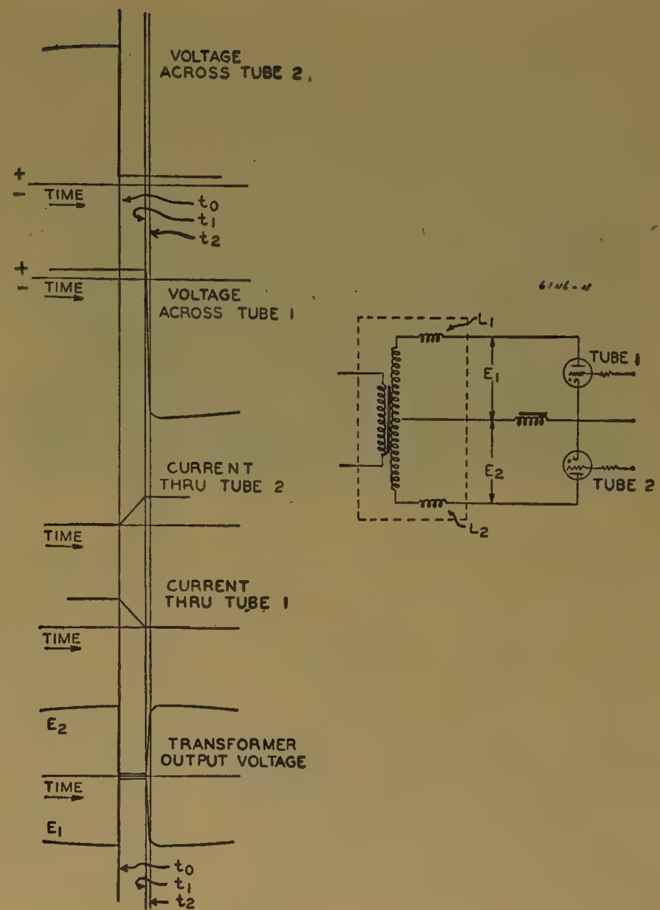

Figure 4. Details of commutation in circuit shown in Figure 3

Showing equivalent circuit and oscillograms of tube voltages, tube currents, and transformer output voltages during current commutation and rise of initial inverse voltage

Table I. Life Test on Tube

Tube Rating: 750 Applied Peak Forward Volts, 5 Amperes Average Current, 60 Amperes Peak Current

Test	Circuit	Frequency	Applied Peak Forward Voltage	Applied Peak Inverse Voltage	Initial Inverse Voltage	Avg Amp	Peak Amp	Hours Life
1	Figure 1	60	155	155	0	10	90	1,955
2	Figure 2	60	15	650	650	5	14	269
3	Figure 3	60	650	650	650	5	10	25
4	Figure 3	60	310	310	310	4	8	120
5	Figure 3	500	680	680	680	5	10	3

Cushion Test Data

Experiments were conducted in which the rate of rise of the initial inverse voltage was modified by adding a capacitor and resistor across each tube, as shown in Figure 5. The capacitor-resistor circuits have been called cushion circuits to indicate that they cushion the fall of anode potential.

Life tests using tubes of lower current rating than those reported in Table I were run under the conditions listed in Table II.

Cleanup occurred in all tubes tested in test 6, whereas none was detectable in any tube after failure in test 7. As before, the particular tube type had been thoroughly tested on various resistive loads and found to be free from cleanup. In addition to demonstrating the effectiveness of cushioning, this test gave further evidence that initial inverse rather than build-up of ionization was responsible for cleanup, as test 7 built up conduction current slightly faster than test 6.

Operation of Cushion

The operation of the cushioned circuit of Figure 5 may be described as follows. For the half cycle prior to entry of tube 2, cushion capacitor C_1 is discharging through R_1 and tube 1. It stands discharged at the moment t_0 when entering tube 2 fires. Current commutation is instituted at t_0 as previously described. Also, entering cushion capacitor C_2 is discharged through R_2 and tube 2. Time t_1 marks the end of current com-

mutation. Thereafter, the initial inverse voltage cannot appear on leaving tube 1 except as current builds up through L_1 and L_2 to cause voltage drops across R_1 and C_1. By proper choice of resistor and capacitor relative to the leakage inductances L_1 and L_2, this rise of initial inverse voltage can be delayed a few microseconds and yet have enough resistance to prevent oscillations. Rates of rise of the order of 10,000,000 volts per second are often satisfactory.

It may help to attempt to visualize the phenomenon occurring within the tube during the commutation and application periods. While conducting current, ionized atoms of xenon populate the electron stream, neutralizing electron space charge. The number of ions varies as some function of current. On a decrease of current, the ion requirements will decrease, allowing the excess to seek electrons and return to atoms. A short time, that is, the order of ten^{-5} seconds, is required for the ionization to readjust to the new conditions. Thus, at the instant that current reaches zero, there still will be present residual ionization representing current flow some microseconds earlier.

At the cessation of current, the initial inverse voltage appears on the anode at the rate established by the circuit. The negative anode attracts residual positively charged ions to it. If ions, being relatively heavy particles, fall through a high potential drop, they acquire enough momentum to sputter or evaporate anode material when they impinge.[2] Sputtering appears to trap gas atoms both in the anode and in the deposited film of sputtered material. If negative voltage is applied slowly enough to allow sweeping out the residual ions at relatively low voltage, sputtering is avoided and cleanup eliminated.

To expedite testing, sensitive pressure gauges were attached to tubes which were then operated under varying current commutation rates and initial inverse voltage application rates. Cleanup could be detected in shorter testing time by this expedient. These data are necessarily tentative, but they do disclose that the permissible rate of rise of initial inverse voltage, above which cleanup sets in, depends upon current commutation conditions. Low reactance transformers, which commutate swiftly, require a lower rate of application of initial inverse voltage, that is, larger cushion capacitors. High reactance transformers may be completely cushioned with quite small capacitances.

It might be noted that, during the course of the experiments using extremely high commutation and recovery rates, there were several cases where the inverse rating of the tube could be changed by adjusting the recovery rates. This undoubtedly has a bearing on the problem of arc-backs, particularly in view of the reported fact that many arc-backs in mercury-arc rectifiers occur just after current commutation.[3,4] This seems to merit a separate inquiry and report.

Based on the pressure gauge tests, the safe rating for the maximum rate of rise of initial inverse voltage is being specified for the present on some tubes as

$\frac{K}{A_{10}}$ = maximum safe rate of rise of intial inverse voltage in millions of volts per second

where

K is a constant for each tube type (6.6 for a typical 6.4 A tube)

A_{10} is the current flowing 10 microseconds before current zero

A_{10} can be determined from measurement of total current flowing and duration of current commutation.

This formulation assumes a linear relation between residual ionization and A_{10} which admittedly is a first approximation, but appears to cover the range of measurable voltage rates moderately well. The factor K usually is larger the smaller the rating of the tube, as might be expected from the smaller ionized volumes. This, plus the fact that fractional ampere circuits do not often have high initial inverse voltage application rates, usually makes it unnecessary to cushion low power tubes.

Table II. Life Test on Tube

Tube Rating: 450 Applied Peak Forward Volts, 1 Ampere Average, 8 Amperes Peak Current

Test	Circuit	Frequency	Applied Peak Forward Volts	Initial Inverse Voltage	Avg Amp	Peak Amp	Hours Life
6	Figure 3	60	423	423	1	2	449
7	Figure 5	60	423	423	1	2	14,200

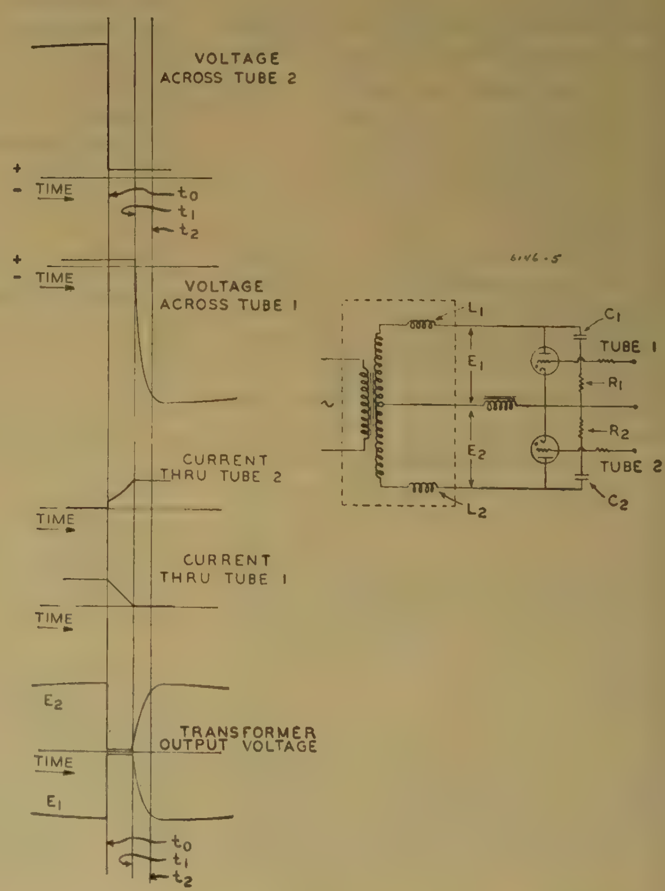

Figure 5. Details of commutation of cushioned circuit

Showing equivalent circuit and oscillograms of tube voltages, tube currents, and transformer output voltages

There is a small power loss in the cushion resistors. It is usually less than two per cent of full output and may be made considerably smaller by proper transformer design.

Measurement

Commutation and cushioning occur so rapidly that they ordinarily escape detection. However, by proper cathode-ray oscilloscope technique they may be observed and measured readily. Anode-to-anode voltage properly attenuated is applied directly to the oscilloscope deflection plates. This voltage wave is viewed with an expanded sweep voltage that is a multiple of supply frequency. It is convenient to apply a synchronizing signal from a small capacitance connected to an anode. The scene thus obtained permits the time of current commutation and the rate of rise of the initial reverse voltage to be scaled off.

There are two sources of error that deserve mention. The signal attenuator must be noninductive and without capacitance errors. The stray capacitance within the oscilloscope between its power source and the oscilloscope ground lead must not be allowed to distort the reading.

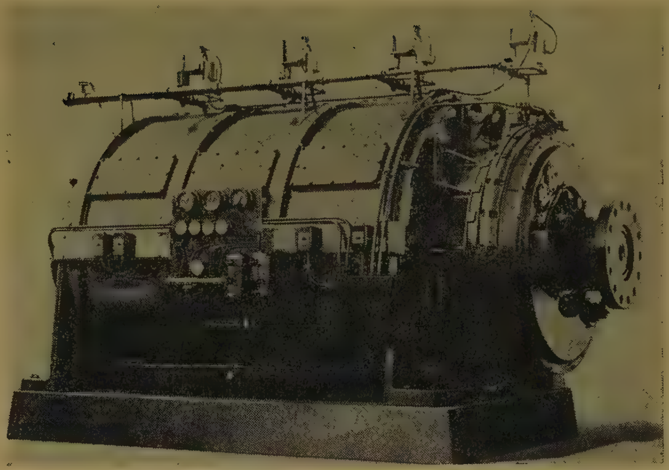

Figure 6 (right). A 12,000 horsepower speed-controlling magnetic clutch regulated by cushioned rectifier

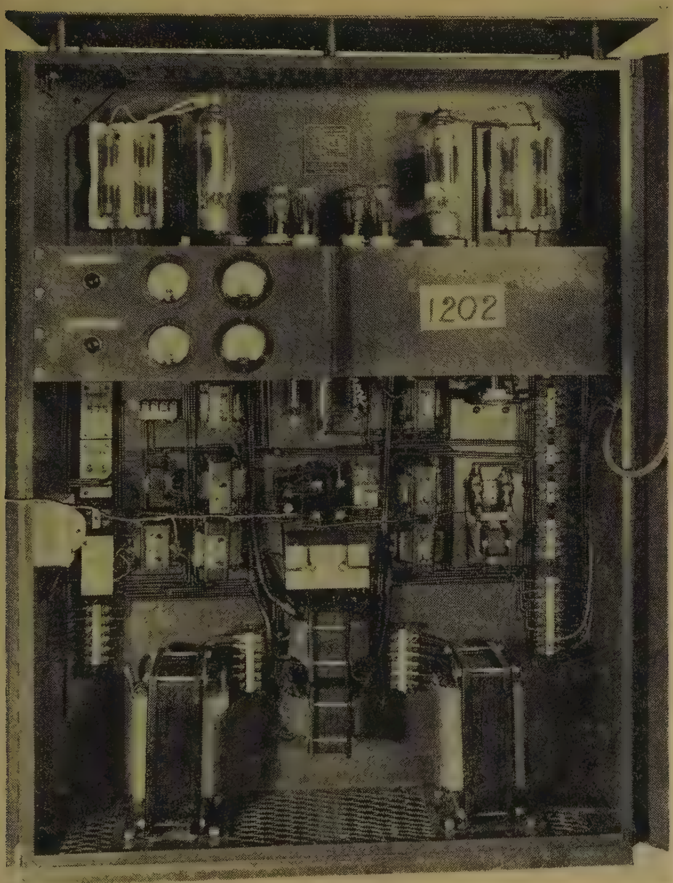

Figure 7 (left). Control rectifier unit for magnetic clutch

Application

The following circuits apply initial inverse voltage if firing is throttled by grid action.

Cushioning is indicated to accommodate the maximum amount of throttling expected in service.

1. Full wave or polyphase rectifiers with inductive load.
2. Polyphase rectifiers with any load operating in continuous current region.
3. Parallel inverters.
4. Half wave control rectifiers with inductive load and back rectifier (only back rectifier requires cushioning).

The conclusions arrived at by life tests have been verified by a large number of actual installations ranging from small controlled rectifiers supplying motor and generator field excitation all the way up to the 12,000-horsepower adjustable speed magnetic clutch shown in Figure 6. Magnetizing power for such units is controlled precisely by a cushioned rectifier as shown in Figure 7. Cushioned control rectifiers have been used successfully over the past two years in many such applications which formerly were impractical.

Conclusions

Experiment indicates that there is a safe rate of rise of initial inverse voltage on gas-filled grid-controlled rectifiers above which gas cleanup may be expected and below which cleanup may be completely avoided.

Tests indicate that the safe rate of rise of initial inverse voltage is a function of the current commutation and of tube construction. This rate of rise may be controlled by proper cushion circuit design.

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Short-Time Current Ratings for Aircraft Wire and Cable

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Synopsis: This paper gives the results of a series of tests on two sizes (AN-16 and AN-8) of insulated aircraft wires and cables, which were made to determine their permissible short-time current-carrying capacities. The times under consideration ranged from 0.01 to 1,000 seconds. A description is given of a very simple but adequate electrical test together with an accompanying insulation test to determine what are considered permissible transient temperatures. Tentative recommendations of short-time current-carrying capacity for aircraft cable sizes AN-22 to AN-00 are made.

WIRE SIZES for aircraft power supply and control circuits, selected on the basis of steady-state temperature rise may prove inadequate for other conditions of operation. Short circuits arising from gunfire or other causes must be anticipated and protected. A great many motors are used to drive the many devices and controls found on the modern plane.

If small wire sizes such as numbers 22 to 6 are used for supplying power to induction motors, d-c motors, and similar loads which may take six times rated current or more during the starting period, and if this branch line is protected by means of temperature overload relays, these branch lines should be capable of carrying six times motor current without exceeding a permissible temperature for a duration corresponding to the time required for the overload relay to trip. It has been recognized commonly that these smaller wire sizes generally will be deficient in their ability to carry this overload for the required period when the selection of the wire size is based entirely on its continuous rating. This deficiency generally is not present in the larger wire sizes.

The purpose of this paper is to supply needed data on aircraft wires to application engineers who sometimes have made the error of using too small a wire size on loads of this character because adequate information generally has not been available. It should also be pointed out that although these data were taken on wires which had insulation appropriate for 30-volt d-c aircraft application, some spot checks were made on a few sizes of wires insulated for 600-volt service and it was found that their short time ratings were quite close to those of the aircraft wires. The insulation on the aircraft wire is actually comparable to that on the 600-volt wire in spite of the lower voltage of the aircraft service. It is believed permissible to apply the present data on aircraft to industrial 600-volt insulated wires, or similar, until data are made available for each specific type of wire. Any difference that may be found between these two types of wires will be on the safe side.

Transient currents in excess of the normal continuous current ratings of the supply wiring and cabling of military aircraft electrical systems are caused by:

1. Starting currents of connected electrical devices when starting under normal conditions.

2. Starting currents of connected electrical devices during abnormal conditions, such as after a period of low temperature shut down during which bearing lubricating oils and greases have congealed or during which connected moving parts have iced or frozen.

3. Actual short circuits resulting from faulted conductors caused by sustained abrasion, vibration, or gunfire.

Circuit breakers or contactors controlled by temperature overload relays are the commonly used means to limit the duration of these currents to prevent the wire temperature from rising to values which would result in damage to the insulating covering of the wires.

The safe time duration depends on the following factors in addition to the current:

1. Initial wire temperature.

2. The heat storage capacity of the copper conductor itself.

3. The heat storage capacity of the wire insulation and its relation to that of the copper conductor.

4. The radiation characteristics of the wire covering.

5. The external air temperature, pressure, and the degree of cooling provided by convection and ventilating air currents.

6. The groupings of the wires in bundles or runs.

The steady-state continuous rating of the wire based on 100 degrees centigrade maximum insulation temperature is governed by radiation characteristics and external cooling factors. These are well established for sea-level conditions in still air, but further work is necessary to study altitude conditions and the special ventilation conditions that exist in a plane in flight. They will not be discussed further in this paper.

The short-circuit momentary-overload condition involves so short a time interval that its safe time duration is governed largely by the initial temperature of the wire and its heat storage capacity at the temperature which may be determined as safe for that time exposure. Values of safe times can be calculated. This paper will discuss these limitations only briefly. For the present, it is assumed that the maximum safe instantaneous copper temperature for all wire sizes is 150 degrees centigrade, for time intervals below one second. Fuse and circuit-breaker designers for many years have used so-called I^2t curves in providing overload protection to circuits. These curves are calculated on the assumption that all the energy is absorbed in heating the copper conductor. They include curves of temperature rise for various currents and times. Such a set of curves is shown on Figure 3. These curves are very satisfactory for periods of time below one second, which include most true short-circuit occurrences.

The problem, then, of selecting wire sizes resolves itself into

1. The selection of wire size for steady-state conditions generally is made on the basis of the continuous rated current capacity of the wire. These ratings are based on rules well established in the industry, though for aircraft use they are based on 100 degrees centigrade maximum operating temperature (70 degrees centigrade rise over 30 degrees centigrade ambient temperature) rather than more usual conservative values.

2. The selection of wire size for overloads of known or estimable duration below one second can be made on the basis of the previously mentioned I^2t curves. It is recommended that 150 degrees centigrade be the maximum allowable temperature to be attained by the copper.

It is recommended that the selection of wire sizes for overloads whose duration is between 1 second and 100 seconds be made on the basis of curves presented in this paper. These curves are given in Figure 4, and are derived from the present series

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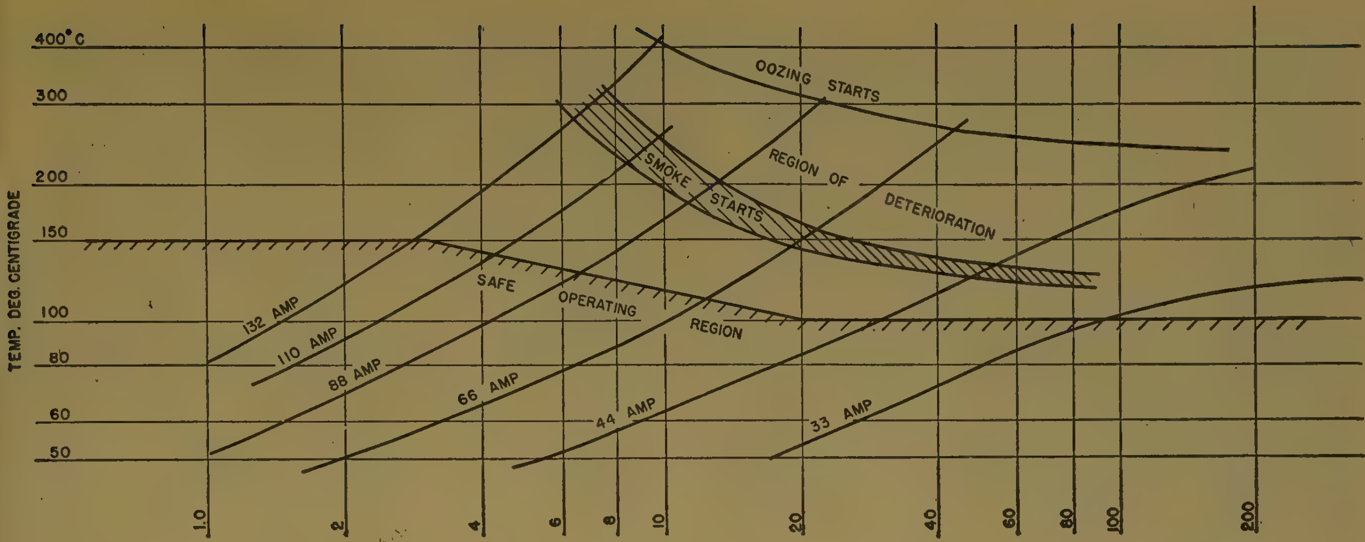


Figure 1. Current-time-temperature curves from 30 degrees centigrade ambient temperature for aircraft cable size AN-16

of tests conducted to study the damaging effects on the insulation of overload currents whose duration was between 0.1 second and several hundred seconds. The tests and the derivation of the curves are described in the following paragraphs.

Method of Test

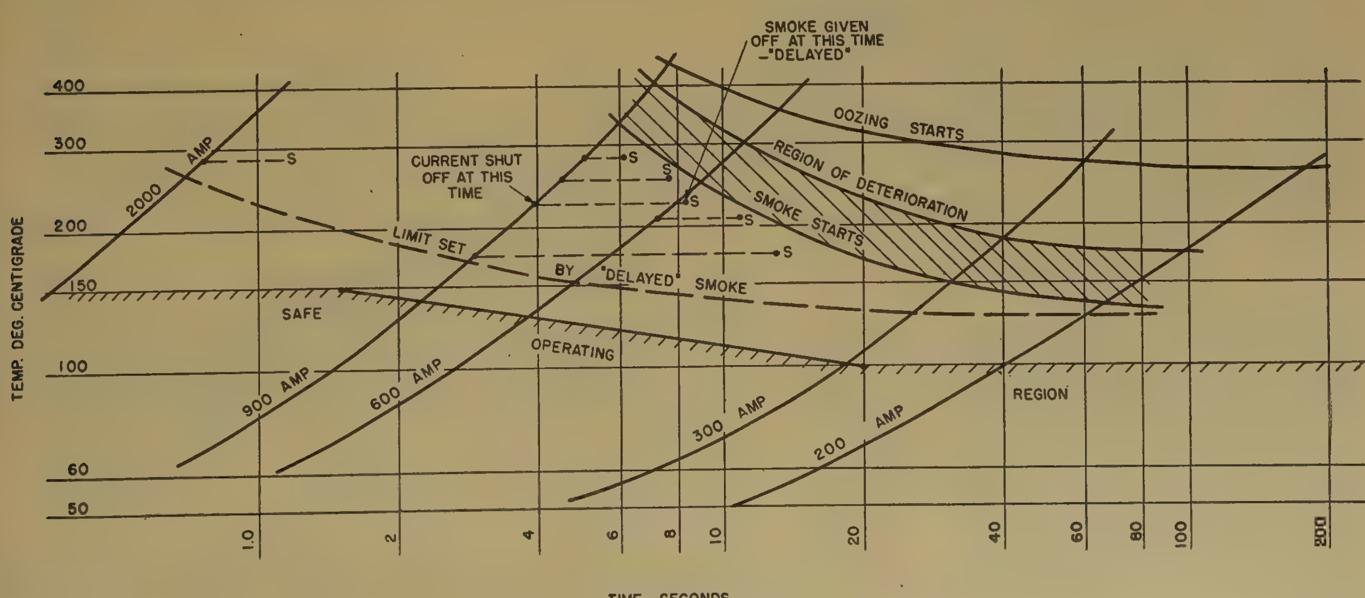
There is no accepted standard method for determining the short-time current-rating of wires. Some engineers have measured temperature rise of wires by means of thermocouples embedded in the center of a wire or brazed to the copper strands. Others have used an oscillograph to obtain a simultaneous reading of current and voltage, while others have used the voltmeter-ammeter method. In this investigation the authors used all three methods and compared their results. Based on this work we will recommend a method which we found was the most practical and yet gave a sufficiently accurate result.

There are two important factors that must be accurately determined when making a test of this kind to ascertain the permissible short-time current-carrying capacities of various wire sizes.

The first is to be sure that the recorded temperatures are accurate, so that the insulation is actually subjected to contact with a wire of the recorded temperature. To illustrate, it was found that the thermocouples gave readings very appreciably lower than the actual wire temperatures as determined by resistance methods. Thermocouples are not suitable for determining temperatures of rapidly heated wires (there are also other limitations for the use of thermocouples).

The second factor is to judge correctly the degree of damage to the insulation that resulted from the above temperature and times. There are standardized methods for measuring the dielectric breakdown and also the cold bend test, but with these having been made there still remains a fair degree of judgment that must be used which should be seasoned with a fair degree of background on this subject. Figures 6A, B, and C, show what is to be expected of such tests. However, as it was found actually that the evolution of smoke was observed even before actual deterioration occurred, and since smoking cannot be tolerated in air-

Figure 2. Current-time-temperature curves from 30 degrees centigrade ambient temperature for aircraft cables size AN-8



planes, its occurrence became the only necessary observation.

A 25-foot length of the wire to be tested was cut and its cold resistance was measured accurately by means of a Kelvin doublebridge in the same ambient temperature where the test was made. A ten-foot central section of this wire was measured accurately for length and marked. This wire then was mounted horizontally about five feet above the test floor and its ends connected to the power supply cables. In order to check the effect of thermoplastic flow of the insulation during the heating, a one-pound weight was suspended from the wire by a hook of one-inch radius made of 1/8-inch diameter steel wire. This weight was located three feet from one end of the wire as indicated in Figure 5. Two steel needles were forced through the insulation and the stranded wire at the two points which marked the ten-foot length. To each of these steel needles was soldered a six-inch length of ten-mil diameter nichrome wire, and this in turn, was soldered to the necessary length of copper wire which ran to the voltmeter terminal. The purpose of the nichrome wire was to minimize the amount of heat that was conducted from the test wire into the voltmeter leads. The circuit arrangement was that shown in Figure 5.

A field controlled d-c generator was the source of power. Current was applied by closing a contactor at the zero time and held constant until the voltage on the voltmeter reached a previously calculated value, corresponding to the temperature to which it was desired to heat the wire. The cold resistance of the 10-foot span was assumed to be 40 per cent of the 25-foot span which had previously been measured by the Kelvin double bridge. The ammeter and voltmeter could be read quite accurately, and the interval of time between the start and the stop also could be timed easily and accurately by stopwatch. Oscillograph records were taken simultaneously of the current, voltage, and time. These provided very accurate measure of all quantities and were in close agreement with the voltmeter-ammeter values. Therefore, since the voltmeter-ammeter method is the simpler, this should be the preferred method, except for time intervals below five seconds where the oscillograph must be used.

Samples of aircraft cable sizes AN-16 and AN-8 were subjected to a number of current values, and at each of these different values of current, wires were allowed to rise to several predetermined values of temperatures so that we could determine the value that should not be exceeded. The rate of temperature rise varied over wide limits.

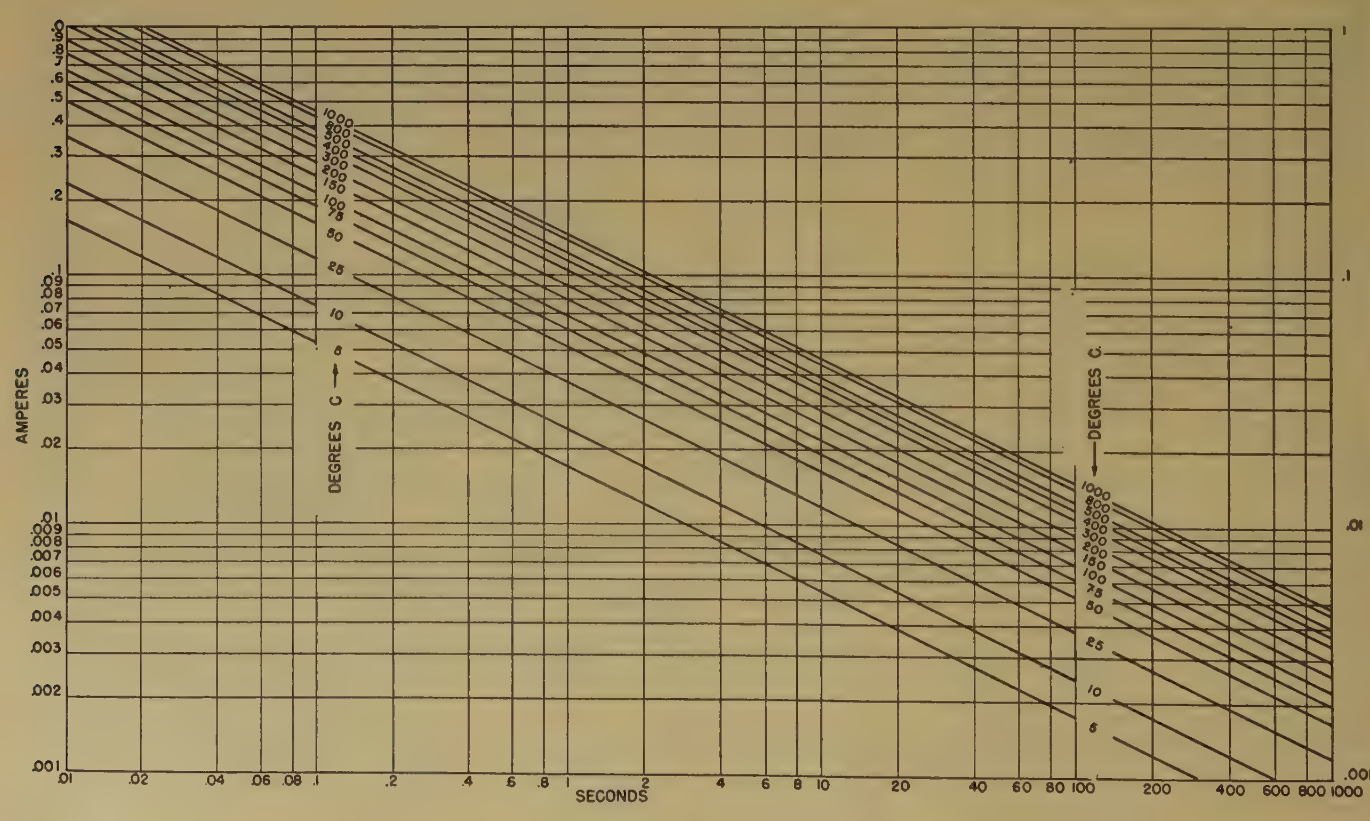
The samples that were allowed to exceed the safe temperatures to a small degree, gave evidence of this by smoking, and other samples that appreciably exceeded this safe temperature demonstrated this by the sudden oozing of the compound out through the insulation surface. The time at which smoking started was observed, and similarly, when the test was continuing until oozing occurred, this time was recorded.

Current-time-temperature curves obtained as a result of these tests were plotted for both sizes of wire, see Figure 1 (size AN-16) and Figure 2 (size AN-8). These curves show temperatures plotted against time for all of the current values tested. The times at which smoke occurred for each current value are shown by the shaded area. The time at which oozing occurred is shown by the wave labeled oozing starts. In the case of the larger wire AN-8, it was observed during the tests that smoking for some values of current occurred after the current had been shut off. This is due to the great relative heat storage capacity in the larger conductor compared to that of its insulation. The smaller wire gives up its heat to its insulation with a rapid drop in temperature. This delayed smoke is shown on Figure 2 and indicated by the notes calling

Figure 3. Time–current curves of one circular mil of annealed copper

Calculated on these assumptions:

1. Radiation neglected
2. Resistance of one centimeter cube of copper at 0 degrees centigrade = 1.589 microhms
3. Temperature coefficient of copper at 0 degrees centigrade = 1/234



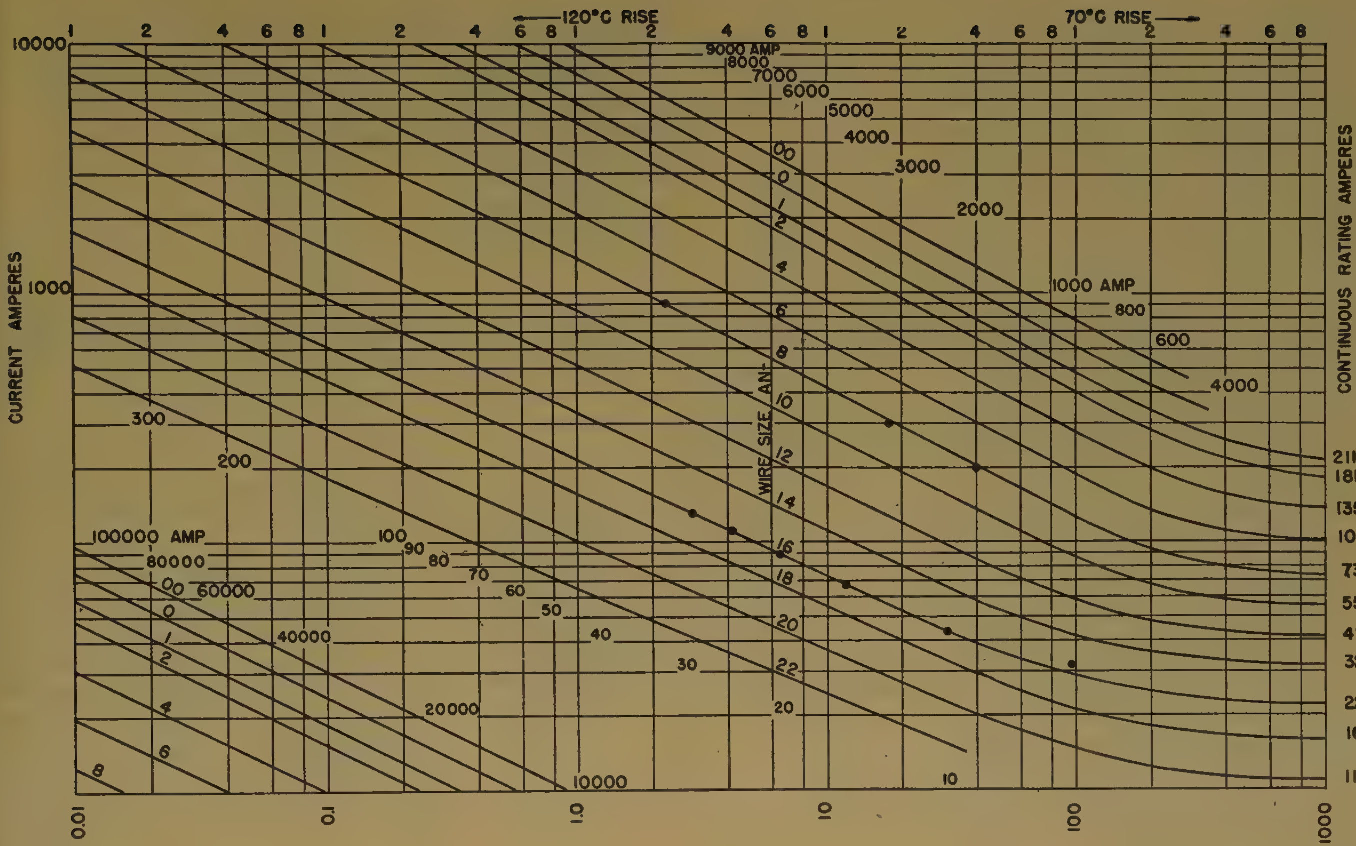


Figure 4. Suggested current–time ratings for aircraft cable AN-J-C-48a for 30 degrees centigrade ambient temperature

attention to this phenomenon. The safe operating region shown on Figures 1 and 2 is drawn at what is considered a sufficiently safe margin below the temperature time limitations imposed by smoke to represent safe operation.

After the test exposure described earlier, each sample of wire was tested for insulation condition as follows:

1. The end sections which supported the one pound weight were tested for dielectric breakdown in accordance with Army-Navy specification, AN-J-C-48a paragraph F-4b(5), entitled, Dielectric Breakdown.

2. One half of the ten foot test section was tested in accordance with Army-Navy specification, AN-J-C-48a paragraph F-4b(6), entitled, Cold Bend Test.

3. The other half of the ten foot test section was tested in accordance with Army-Navy specification, AN-J-C-48a paragraph F-4b(4)c, entitled, Water Immersion Test.

The results of these dielectric breakdown tests are presented graphically in Figures 6A, B, and C, respectively.

It should be noted that a reasonably large number of tests were made because we were endeavoring to find the short time permissible temperatures. But if the temperatures that are recommended here are accepted, any future tests can be simplified very greatly by merely finding the time required for any selected current to produce the maximum temperature shown in Figures 1 and 2.

Recommended Test Procedure

Establish by trial a current which will produce the allowable temperature rise (usually 70 degrees centigrade) in approximately 20 seconds (between 10 and 40 seconds) and determine by test the exact time required. On a sheet similar to Figure 4, locate this one point of time and current. Then from Figure 3 calculate the current required to heat (on a no-loss basis) the wire size in question to a value equal to 150 degrees centigrade minus the assumed ambient temperature in 0.01 second. This will give another point on the afore-mentioned sheet. Then draw a straight line between these two points and that will be the permissible short-time current-carrying ability of that wire size. By testing in this simple manner about four or five sizes of wire, for any one type of wire the whole range of ratings can be drawn in by proportioning the intermediate sizes in proportion to their respective cross sections.

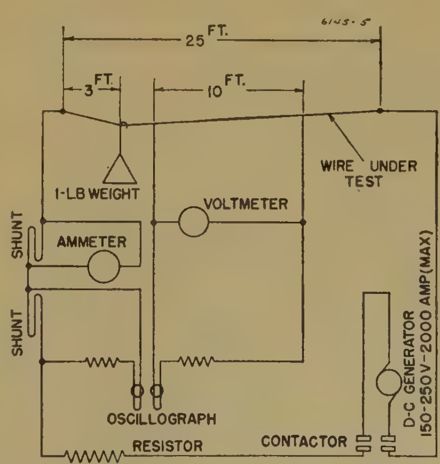


Figure 5. Test circuit for current-time-temperature rise test on aircraft cables

Beyond the 100-second region this straight line can be extended so as to become asymptotic to the current line, corresponding to the continuous rating of this wire, which usually is available for the assumed ambient temperature.

Discussion of Results

The smoke criterion of damage occurred before any other evidence of deterioration. On the larger wire, size AN-8, smoke often was delayed considerably after the current had been removed. This is indicated by the delayed smoke limit curve, Figure 2. This is evidence of the large energy storage in the conductor during the short circuit, and the low heat capacity and heat transfer constant of the insulation compared to that of the large conductor.

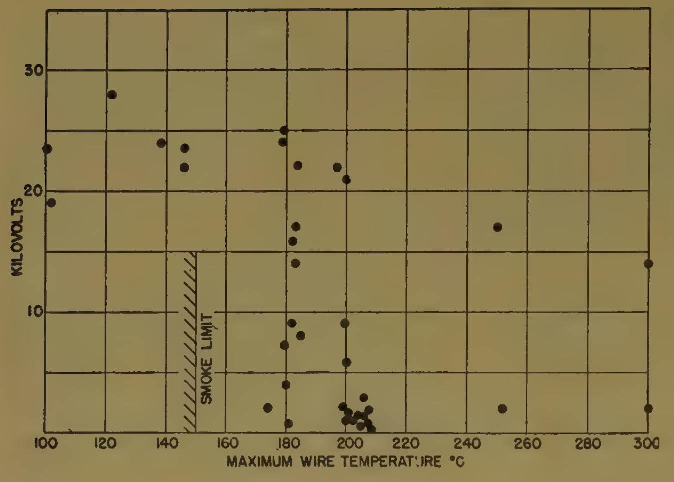


Figure 6A. Kilovolts breakdown in water at 25 degrees centigrade of samples containing point of attachment of one pound weight, one 1/8-inch diameter wire hook, AN-J-C-48a paragraph F-4b(5)

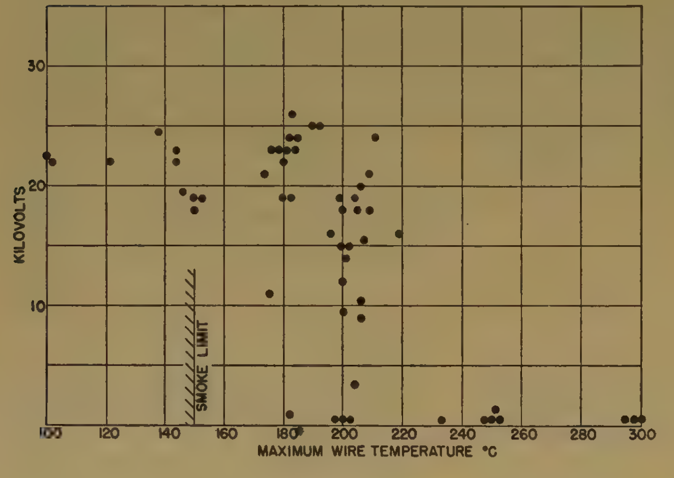


Figure 6B. Cold bend test kilovolts breakdown of samples in water after one hour at 40 degrees centigrade, AN-J-C-48a paragraph F-4b(6)

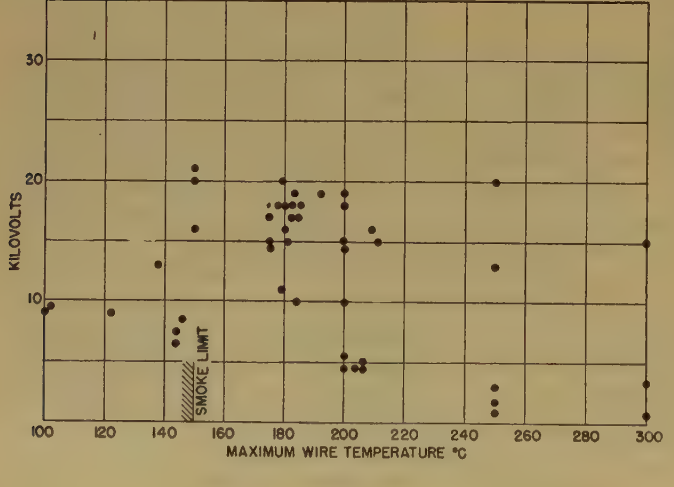


Figure 6C. Kilovolts breakdown in water at 50 degrees centigrade after immersion in water for 20 hours at 50 degrees centigrade AN-J-C-48a paragraph F-4b(4)c

The dielectric breakdown values given in Figures 6A, B, and C, gave evidence of deterioration only when the insulation had been carried to a considerably higher temperature than that at which smoke first appeared. Dielectric breakdown results vary over a large range, but do indicate definite deterioration when the wire has been exposed to temperatures of 200 degrees centigrade for the short time intervals covered by these tests.

The appearance of smoke, then, is the first indication of abnormal conditions, and if current is not removed immediately a condition of serious damage soon will result. It is also a condition which cannot be tolerated for psychological reasons in airplanes. The instinctive association of smoke with impending or possible fire is so strong, and the hazard of fire so great in an airplane, that we have adopted the slightest appearance of smoke as the limiting condition of current and time below which we must operate.

On each of Figures 1 and 2 is shown a so-called safe operation region which has been arbitrarily bounded as follows:

Wire, Size	Time, Seconds	Maximum Temperature, Deg C
AN-8	0–1.5	150
AN-8	Over 20	100
AN-8	1.5–20	From sloping line connecting these temperatures
AN-16	0–2.5	150
AN-16	Over 20	100
AN-16	2.5–20	From sloping line connecting these temperatures

Using this boundary of the safe operating region, values of times for the various test currents were obtained for these two wire sizes, in the time range 1 to 100 seconds. These are plotted on Figure 4. To extend these curves to the left to lower times, values of current to produce 150 degrees centigrade at the end of 0.01 second were obtained from Figure 3, namely, 12,300 amperes and 1,700 amperes for sizes AN-8 and AN-16, respectively. These points were plotted and connected by straight lines to the curves already drawn between 1 and 100 seconds.

To extend the curves to the right for times greater than 100 seconds, the curves were connected by smooth arcs to the steady-state values, 73 amperes and 22 amperes, respectively, which were obtained from the steady-state ratings given by the AIEE committee on aircraft wire and cable.

Curves for other wire sizes between 16 and 8 were interpolated on the basis of the theoretical spacing. Similar extrapolation was made to wire sizes 20 and 00. These curves are tentative only and subject to verification by further tests.

Conclusion and Recommendations

Current-time-temperature tests on wires, carried to the point at which smoke first appears, provide a suitable method of determining the short-time current-carrying characteristics of wires. The temperature-time measurements must be made carefully and accurately. Wire temperature should be determined by copper resistance, using a voltmeter-ammeter method for times down to five seconds and using an oscillograph method for shorter times. Thermocouple measurements are not recommended.

The suggested current time ratings for aircraft cable AN-J-C-48a presented in Figure 4 based on a 30 degrees centigrade ambient temperature is recommended for consideration.

Low-Current-Secondary Current Transformers

AN AIEE COMMITTEE REPORT

APPLICATION REQUIREMENTS for current transformers sometimes indicate the use of secondary currents lower than the generally accepted standard 5-ampere rating. Very little published information is available on this subject, other than a description of a few specific installations where such transformers have been applied. The natural question arises as to the advantages and disadvantages resulting from the use of current transformers with secondary ratings below five amperes, and under what circumstances their use should be considered. Because many of the applications of low-current secondaries involved relay protection, the AIEE relay subcommittee appointed a working group in October 1943 to study the problem, and from that study, gives recommended practice for their use.

Applications

The principal application of low-current-secondary current transformers is in those installations involving long runs of secondary leads where the resistance of the leads (utilizing normal size copper conductors) becomes very high, thus imposing on the current transformer excessive lead burden in comparison to the useful burden. A brief summary of the problems to be considered for such installations is covered in this paper.

Another application of current transformers using secondary ratings of other than five amperes, occurs in those installations requiring the paralleling of the secondaries of transformers subject to widely different normal primary load currents. In such cases, the secondary rating is primarily one of convenience and economy. The secondary circuit constants, transformer turn ratios and performance, usually are fixed on a 5-ampere secondary basis. This is dictated by transformers in some of the circuits involved having the standard 5-ampere secondary rating.

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This report was prepared by a working group of the relay subcommittee, AIEE committee on protective devices, consisting of S. C. Leyland (M '43, *sponsor*), manager, relay section, meter engineering department, Westinghouse Electric Corporation, Newark, N. J.; F. C. Poage (M '38, *former sponsor*), electrical engineer, engineering department, Ebasco Services, Inc., New York, N. Y.; R. E. Neidig (A '38), relay engineer, Metropolitan Edison Company, Reading, Pa.; L. F. Kennedy (M '39), relay application engineer, General Electric Company, Schenectady, N. Y.; E. C. Schurch, (M '35), senior engineer, electrical department, United States Bureau of Reclamation, Denver, Colo.

Comparison of Solutions

The use of current transformers with secondary ratings below five amperes for the above applications has certain advantages and disadvantages which must be carefully considered in evaluating the gains resulting in the reduction of effective burden of the secondary leads. Any general comparison of these advantages and disadvantages should be made on the common basis of equal volt-ampere energy requirements for the relays, instruments, and meters, irrespective of the secondary rating. Changing the current rating of these devices in no way reduces the energy required for their operation. A reduction in energy is just as desirable on the basis of five amperes as on a lower current rating basis, and minimum operating energy is a controlling factor in the design of this type of equipment.

On the basis of constant volt-ampere energy requirements, a reduction in secondary current rating results in a corresponding increase in secondary voltage rating. The energy consumption in the external devices and the internal losses in the transformer are not materially affected, thus, the only gain is the reduction in energy consumed by the secondary leads between the transformer and external burden.

As an example, if it is assumed that the distance between the current transformer and the instruments or relays is 500 feet, and standard control cable such as one consisting of conductors each having 19 strands of number 22 wire is used, the burden imposed on the transformer by each secondary lead is 11 watts at 5 amperes. Under the same conditions, with a one-ampere secondary, the lead burden would be 1/25 or 0.44 watt per lead at rated current. Thus, in some cases, smaller wire could be used with a resulting saving of copper. Such savings may be limited, however, by the necessity of retaining a larger wire size to obtain mechanical strength. Unless the lead burden is a fair portion of the total burden, this reduction in lead burden materially does not lessen the performance requirements of the current transformer.

In evaluating the benefits gained by the reduction in effective secondary lead burden, consideration must also be given to the effects on the insulation requirements of the secondary circuit, and the availability of standard devices suitable for use on the lower current rating basis.

With low-current secondary currents and the corresponding increase in secondary voltage (assuming equal energy requirements), due consideration must be given to the required increase in secondary voltage. For instance, the present "American Standards for Instrument Transformers" recognize a maximum relaying accuracy classification of 800 volts on the secondary terminals at 100 amperes (20 times rated) secondary current, based on five amperes. On a one-ampere basis, the secondary terminal voltage would be 4,000 volts, and the secondary insulation should be well in excess of this value for safe operation. Furthermore, the present open-circuit voltages on switchboard transformers commonly run from 500 to 5,000 volts crest. The logical basis for consideration of a one-ampere-secondary, therefore, is that the open-circuit voltages will be five times as great, or from 2,500 to 25,000 volts crest. It is apparent that the standard one-minute potential test of 2,500 volts to ground for the secondaries of current transformers should be increased to 12,500 volts for one-ampere secondary ratings. Similarly, relays and instruments used with one-ampere secondary transformers should have the one-minute potential test of 1,500 volts to ground increased to 7,500 volts.

The above requirements for insulation are based on general applications. However, for those applications requiring the use of low-current-secondary current transformers to minimize lead burden, and which will be subjected to considerably less than 20 times normal current, a suitable transformer could be made with an insulation level to meet the actual secondary voltage requirements.

While it is quite true that the secondary voltages existing at normal currents are low, the high voltages occurring on high values of overcurrents and on open circuits sometimes are overlooked. In cases

where only meters or instruments are involved, secondary overvoltage protection may be applied without too much complication. This is particularly true where such protective devices can be permitted to short-circuit the transformer on excessive overvoltage. However, overvoltage protective devices which short-circuit the transformer secondary cannot be used where protective relays are involved, because incorrect relay operation would result. For the foregoing reasons, it is sometimes feasible when considering less than 5-ampere secondary ratings to go to a lower secondary current rating when only meters and instruments are used. When protective relays are used, the problem requires considerable more study.

When considering the use of low secondary current ratings, attention must be given to the use of relays having suitable current ratings. The present complete line of relays (some available with low current taps) was developed to cover the requirements of the field on the basis of the standard 5-ampere secondary transformer ratings. Therefore, their equivalent on a 1-ampere basis does not exist. While a limited number of relays have coils suitable for certain low-current rating applications, to obtain a complete line of such devices with suitably co-ordinated taps and insulation levels would require major development.

In the applications where there is a definite need for a lower secondary current resulting from excessive length of secondary leads, it is desirable to retain standard 5-ampere rating relays and instruments. This may be accomplished by using standard 5-ampere rating main transformers, relays, and instruments with auxiliary current transformers at the main current transformers and the switchboard. It also is possible to use main current transformers with low current secondary rating in conjunction with auxiliary transformers at the switchboard only. It must be remembered, however, that auxiliary current transformers introduce additional burden which tends to offset the gain resulting from reduction in secondary lead burden, and at the same time, lower the accuracy of the over-all transformation because of their own errors. The performance characteristics of the auxiliary current transformers must be as good as those required of the main current transformers, and they should be provided with sufficient insulation to protect the instruments and relays from the higher secondary voltages accompanying the lower currents. Where auxiliary transformers are used, the switchboard circuits should be protected from excessive voltage by the use of suitable short-circuiting devices. Proper consideration must be given to the performance of the auxiliary transformers on a d-c transient having a long time-constant. Auxiliary current transformers fulfilling these requirements increase the cost, require installation space, and reduce the net gain.

Conclusions

In view of the many disadvantages of low current secondaries, compared to the advantage of reduced secondary lead burden or saving in copper, they should not be considered except in special cases. These special cases occur where the secondary lead burden is the major portion of the total burden and definitely causes errors that cannot be tolerated. Then the problem should be studied thoroughly in the light of the foregoing factors. The secondary current rating selected and the secondary circuit arrangements used should be determined as best meets the requirements of the individual application.

The working group feels that the problem of low-current-secondary current transformers is so special that no consideration should be given to standardization at this time.

A Unique Engine-Synchronism Indicator for Aircraft

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ASSOCIATE AIEE

RECENT aircraft design has emphasized multiengine airplanes, especially for commercial transports and military types. Two- and 4-engine planes are commonplace and 6-engine craft are proposed for the near future. Engine synchronism indication, usually between a master engine and each of the others, becomes more important as the number of engines on the airplane increases. It is possible to synchronize two engines by ear, but this method becomes exceedingly difficult for aircraft with more than two engines.

The synchronism indicator, or synchroscope, has been developed to obtain a visual indication of the degree of synchronism that exists between the master and each of the other engines. A discussion of the needs for engine synchronism indication and of the means by which it is obtained, will be presented.

Reasons for Synchroscopes

Perfectly synchronized airplane engines are desirable for several reasons. Better flight efficiency[1] is obtained when the engines are running at the same speed. Planes with constant pitch propellers tend to yaw if the engines are not synchronized exactly. If a condition of near-synchronism exists, such as occurs when synchronizing is attempted with the tachometer indicators alone, a vibration of low frequency is produced. This beat frequency is the arithmetical difference between the high frequency vibrations produced by each engine. Its presence causes high stresses[2] to be set up in the airplane structure which can cause severe damage to the airplane. Fatigue and discomfort for the passengers and crew of the aircraft also are caused by the presence of these vibrations. For example, the Air Forces have found it very desirable to use synchroscopes on transport planes to avoid these harmful vibrations

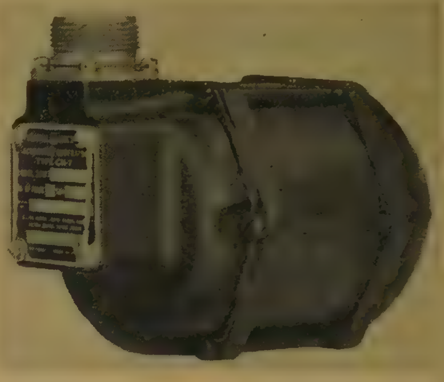

Figure 1. Tachometer generator

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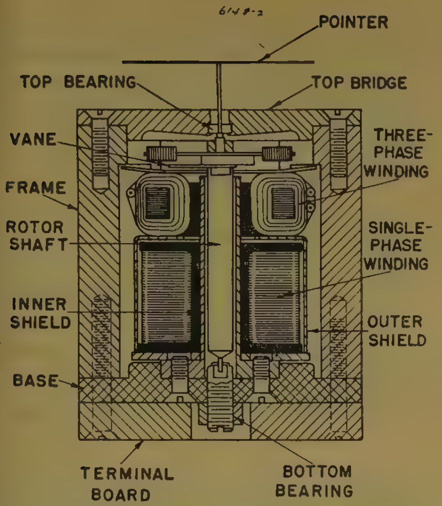


Figure 2. Cross-section view of synchroscope

when carrying sick and wounded personnel.

To promote commercial flying to the greatest degree, passenger comfort is of paramount importance, especially on sleeper-type planes. The synchroscope has been found to be a simple and satisfactory instrument for adjusting engine speeds to obtain efficient flight operation, minimized stresses on the aircraft structure, and to provide a high degree of comfort for the passengers.

Equipment Needed to Indicate Synchronism

No additional source of power or transmitting device is necessary to obtain engine-synchronism indication. Electric tachometer generators[3,4] already installed to operate the tachometer indicators will supply sufficient additional power to operate the synchroscopes. The only additional weight necessary is that of the indicators themselves and the connecting leads.

The most satisfactory tachometer generator has an a-c output with a frequency proportional to engine speed. A typical generator of this kind is illustrated in Figure 1. It is customary to use a reduction gear with a ratio of two to one to drive the generator. Its 4-pole rotor is made of a permanent magnet material such as alnico II. This magnet, as it rotates, induces in the 3-phase winding on the generator stator, a 3-phase a-c voltage having a frequency proportional to the engine speed.

To obtain an indication of the degree of synchronism between two engines, a measurement is made of the frequency difference between the outputs of the generators connected to each engine. To perform this function, the synchroscope uses two stator windings and a rotor. Figure 2 illustrates the mechanical construction of the instrument, while Figure 3 shows the electric circuit. One coil is a 3-phase winding, placed in the form of a toroid on a circular core of high permeability material, such as mu-metal. This winding usually is connected to the generator on the reference, or master engine, which is customarily the one at the extreme left.

Below the 3-phase coil is a single-phase circular stator coil, wound with its axis at

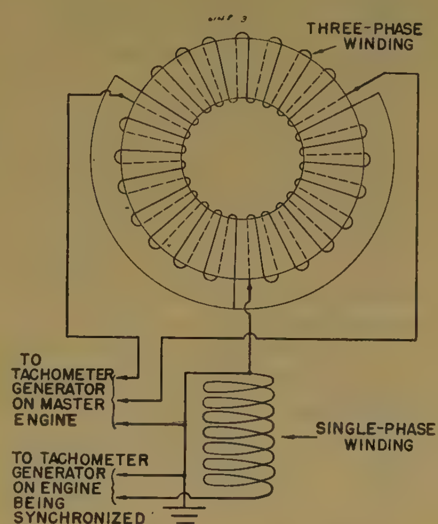


Figure 3. Electric circuit diagram of synchroscope

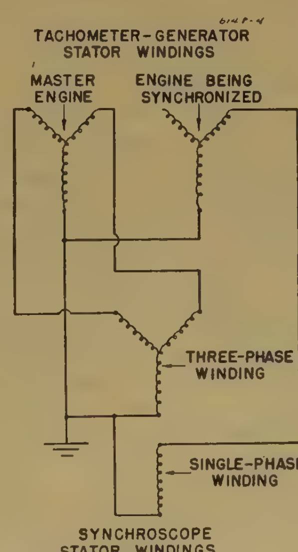


Figure 4. Wiring diagram for synchroscope and tachometer generators

right angles to the plane of the 3-phase winding. This coil is connected to one of the phases of the tachometer generator driven by the engine whose speed is being compared to the master engine.

A rotatable shaft and vane assembly of high permeability material, and supported by suitable bearings, is placed in the instrument so that it can be acted upon by the magnetic fields produced by the two stator windings. A pointer on the shaft gives visual indication of the condition of synchronism that exists between the two engines in question. This construction, embodying two stator coils and a soft-iron rotor, is similar somewhat to that used in switchboard type synchroscopes.[5]

Principle of Operation

When the master engine is running with the synchronism indicator connected as in Figure 4, a rotating magnetic field is produced by the 3-phase winding. This field rotates at the same speed as the engine because the 4-pole tachometer-generator rotor is driven at one-half engine speed. When the engine connected to the single-phase coil is running, there is produced in this winding a magnetic field that alternates at a rate equal to the speed of the connected engine.

As shown in Figure 5, the flux from the single-phase winding crosses the air gap to the shaft and then travels to the vane. After leaving the vane, it passes through the 3-phase core and links with the field from the 3-phase winding. The flux then flows through a washer and a shield of high-permeability material to a second washer at the bottom of the single-phase coil. From the bottom washer the single-phase flux passes to the inner shield, or sleeve, which surrounds the shaft, and then crosses the air gap to the shaft, completing the circuit. All air gaps in this circuit are kept as short as possible to hold the reluctance to a minimum.

When the single-phase winding is energized, its flux, alternating at a frequency proportional to the speed of the engine to

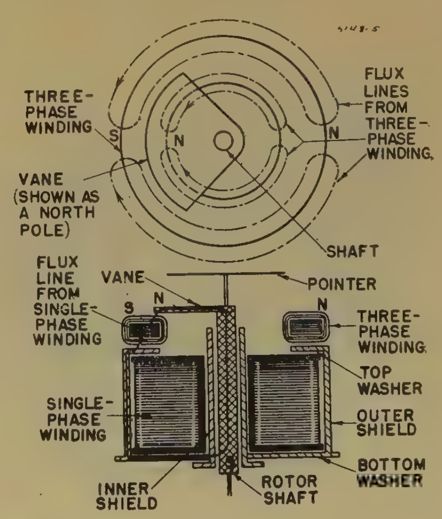


Figure 5. Magnetic circuit diagram of synchroscope

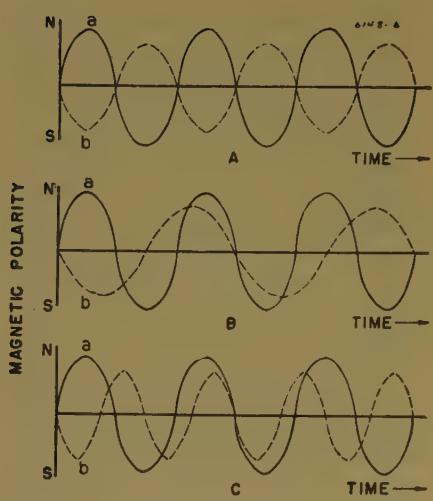


Figure 6. Flux relationships in synchroscope

A—Engines synchronized
B—Second engine running slower than master engine
C—Second engine running faster than master engine
a—Magnetic polarity of a point on the upper surface of the 3-phase winding
b—Magnetic polarity of the rotor vane

Figure 7. Three synchroscope mechanisms in a 17/8-inch diameter case

Figure 8. Five synchroscope mechanisms in a 23/4-inch diameter case

which it is connected, causes the magnetic polarity of the rotor vane to reverse at the same rate. With the field of the 3-phase winding rotating at a rate proportional to the master engine speed, the rotor assumes a position such that when the vane is a north pole, a south pole from the 3-phase field is at the same point. Figure 6*A* shows the flux relationship in the synchroscope when synchronism exists between the two engines. If the flux in the shaft-and-vane assembly alternates at the same rate as the 3-phase field rotates, the synchroscope rotor will remain stationary, as the fluxes from the two fields produce no rotational forces in the rotor vane. As the vane changes from a north to a south pole, the field from the 3-phase winding rotates 180 degrees from its previous position, so that its south pole is replaced by a north pole.

Figure 6*B* illustrates the flux relationship that exists in the instrument when the engine being synchronized is running slower than the master engine. Under these conditions, the rotating field will have moved more than 180 degrees by the time the vane polarity has changed from maximum north to maximum south. The vane-and-rotor assembly will rotate counterclockwise at a speed sufficiently great to make up the difference in frequency between the two fluxes. This enables the two fields to remain locked together. The torque which causes the rotation is obtained from the forces of attraction and repulsion produced by the two fields acting on the rotor vane.

Figure 6*C* shows the relative position of the two fields that exists in the synchroscope when the second engine is running faster than the master engine. The frequency of the single-phase flux is higher than the frequency of the 3-phase flux, and the rotating field will have moved less than 180 degrees when the vane polarity has changed from maximum north to maximum south. The rotor will turn clockwise at a speed great enough to make up for the difference in frequency between the two fluxes.

For the typical arrangement of 4-pole rotors in generators driven at one-half engine speed, the synchroscope rotor will turn at a speed equal to the difference between the speeds of the two engines. For other arrangements, the synchroscope rotor will turn at a speed proportional to the difference between the speeds of the two engines. The pointer at the end of the rotor shaft, therefore, gives a visual indication of the condition of synchronism that exists between the two engines in question. To synchronize an engine with the master engine, all the pilot need do is to note whether it is running too fast or too slow, as indicated by the direction of rotation of the synchroscope pointer and the markings on the instrument dial, and adjust its speed accordingly until the indicator pointer becomes stationary.

Design Features

Panel space is at a high premium in present day aircraft, and it is very desirable and necessary to reduce the amount of panel space occupied by any instrument. For the synchroscope this has been accom-

Figure 9. Disassembled view of synchroscope mechanism showing top of 3-phase winding, top bridge, and rotor assembly

plished by an unusually compact instrument-mechanism design. Figure 7 shows three synchroscope mechanisms in a $1^7/_8$-inch diameter case to provide synchronism indication for 4-engine planes. Figure 8 illustrates five synchroscope mechanisms in a $2^3/_4$-inch diameter case for 6-engine aircraft. If desired, a single mechanism can be used with a selector switch in any multiengine airplane, or it may be incorporated in a dual tachometer indicator.

Simplicity of construction and mechanical sturdiness are desirable features for aircraft instruments. These features are illustrated by Figure 9 which shows a disassembled synchroscope mechanism. Placing both windings on the stator, thus eliminating brushes and all other connections to the rotor, permits the synchroscope to be built strongly and yet simply enough to allow easy disassembly. Less than five minutes are required, using only a small screwdriver, to take apart and reassemble the complete mechanism.

Operating Characteristics

Two primary operating characteristics determine how satisfactory a synchroscope is for aircraft applications.[6] First, the rotor should begin to turn when the differential speed between the two engines being synchronized is 150 rpm or more over an engine speed range of 1,400 to 4,500 rpm. It should continue to rotate

until this differential speed has been reduced to zero. Second, when the rotor is turning, it should rotate smoothly, especially when the differential speed is 5 rpm or less.

Smooth operation is dependent on two characteristics of the synchroscope. The natural period of oscillation of the system, made up of the rotor mass and the magnetic coupling between the vane and the rotating 3-phase field, must be above the maximum speed at which the instrument is to operate. The strength of the 3-phase field must be uniform around the periphery of the 3-phase winding. If these conditions do not exist, erratic operation of the rotor assembly will occur. By keeping the moment of inertia of the rotating system as low as possible, and by using a toroidal construction for the 3-phase core and coil assembly as previously described, smooth operation is obtained.

Figure 10 shows typical starting characteristics of a synchroscope built according to the design previously discussed. Small magnetic irregularities in the instrument add to the starting torque in one direction and subtract from it in the opposite direction. For this reason, the starting characteristics vary slightly, depending on the direction in which the rotor is turning.

Figure 10. Starting characteristics of synchroscope at room temperature

A—Rotor turning clockwise
B—Rotor turning counterclockwise
C—Average of A and B

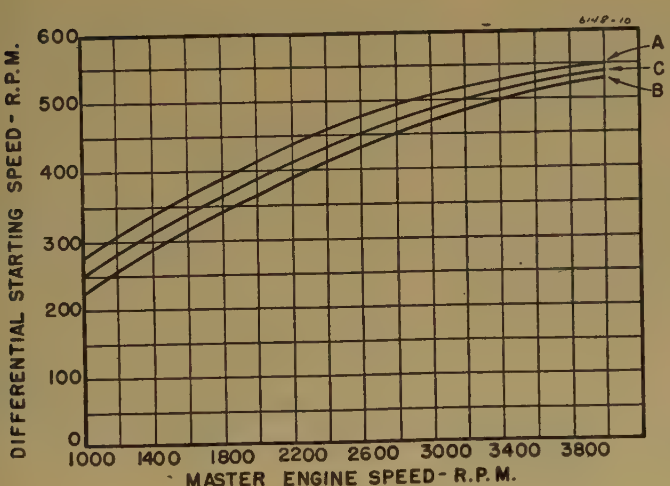


Figure 11 shows the starting characteristics of the synchroscope at high and low ambient temperature. The torque exerted on the rotor vane is an inverse function of the temperature of the synchroscope windings, as the resistance of the coils is a direct function of their temperature.

To keep the windings at a safe operating temperature, and to prevent fogging of the instrument cover glass, the temperature rise in the instrument must be considered. Table I shows these data obtained on a typical synchroscope sample.

Table I

Time, Hours	Room Temp, C	Max Safe Temp Rise, C	Temperature Rise, C: Single Phase Coil	Three Phase Coils: AB	BC	AC
0	25		0	0	0	0
1.5	26	40	23	22	25	18

Experience obtained from other instruments indicates that satisfactory service will be obtained at these operating temperatures.

Application Other Than Aircraft

While it has been shown that the synchroscope has a definite aircraft application, its usefulness does not end there. A synchronism indicator such as has been discussed can be used for many other applications where it is necessary to indicate or control the difference in the speeds of two engines.

If 3-phase power is not available, a conventional split-phase[5] design such as is used with some types of single-phase motors can be utilized to operate the synchroscope.

The instrument also can be used as a power-factor or phase-angle indicator when one winding is excited by a voltage in phase with the supply voltage, and the other winding by a voltage in phase with the load current. The position taken by the pointer will be a function of the phase angle between the voltage and current.

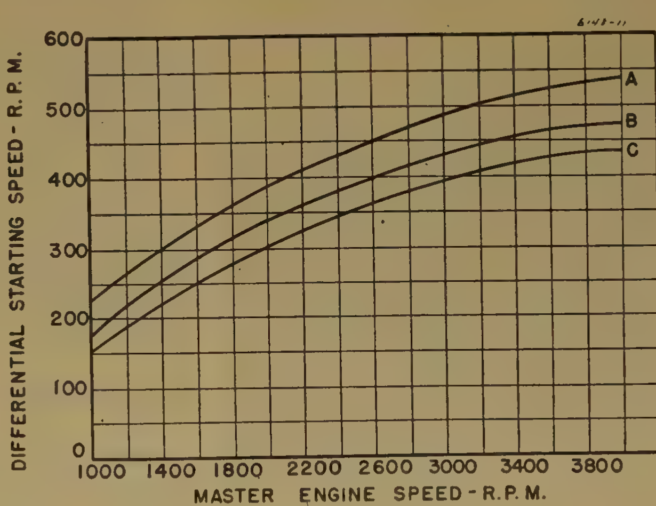
Conclusions

Visual synchronism indication has been shown to contribute appreciably to the satisfactory operation of multiengine aircraft. Improved passenger and crew comfort, lessened vibrational stresses in the aircraft itself, and better flight efficiency are among these contributions.

Characteristics of the synchroscope are simplicity of construction, ease of maintenance, small size, and low weight. These features, coupled with the desirable operating characteristics described previously, make the synchroscope a useful addition to the instrument panels of multiengine airplanes.

Figure 11. Starting characteristics of synchroscope at high and low ambient temperatures

A—Synchroscope at −55 degrees centigrade
B—Synchroscope at +25 degrees centigrade
C—Synchroscope at +71 degrees centigrade



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Nonlinear Commutating Reactors for Rectifiers

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MEMBER AIEE

ONE of the prime objectives in rectifier development during the past 20 years has been the decrease in frequency of arc-back. The factors involved in arc-back have received considerable attention, and it now appears that arc-backs of typical mercury arc rectifiers are caused by the following:

1. Breakdown of the mercury vapor because of inverse voltage.

2. Conditions favorable for cathode spot formation on the anode at the end of the commutation period.

The former cause is not present under normal operating conditions, but may become effective at excessive vapor pressures, as when condensed mercury is permitted to come into contact with a hot anode. The latter cause has been treated in some detail by Kingdon and Lawton.[1]

Phase of Arc-Back

Kingdon has proposed that the cause of arc-back is the charging by positive ions of small insulating particles on the surface of the anode, thereby causing local fields strong enough for runaway field emission. The experiments of Kingdon and Lawton on tubes with 1-centimeter anode-grid spacing indicate that the probability of commutation arc-backs varies with the product of initial inverse voltage and final commutation rate. In a structure with close spacing between the anode and deionizing members, the rate of ion diffusion during commutation is so rapid that the ion density at the end of commutation depends on the rate of change of current at the end of commutation. The number of ions collected by a particle on the anode at the end of commutation determines its resultant potential. This number is proportional to the inverse voltage and the density of ionization.

Hull and Elder have described tests

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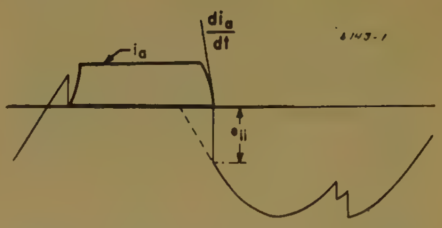

Figure 1. Anode voltage and current in 3-phase rectifier showing elements of rectifier duty

where the factors were controlled so that arc-backs occurred both at commutation and at later times.[2] They concluded that practical rectifiers of correct design will arc-back at the end of commutation rather than some later time, unless conditions are favorable for glow discharge.

There is evidence that the causes of arc-backs in practical rectifiers lead to a distribution of arc-backs throughout the inverse period.[3] A considerable proportion of these occur at the end of commutation. Furthermore, it is common experience that the probability of arc-back at a given load increases with the amount of phase retard and that rectifiers having a rating based on a considerable amount of phase retard can be given a greater rating with no phase retard.

Rectifier Duty

Figure 1 shows the current and voltage in one anode of a 3-phase rectifier and the quantities involved in anode duty. If

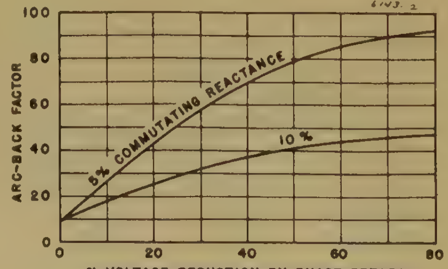

Figure 2. Effect of phase retard and circuit reactance on arc-back factor F_a

$$F_a = e_{ii} \times \frac{di_a}{dt}$$

the effect of residual ionization is denoted by an arc-back factor F_a, then

$$F_a = e_{ii} \times di_a/dt$$
$$= 2\pi f e_{ii}^2/X_c$$

In Kingdon's experiments the frequency of arc-back was approximately proportional to the tenth power of F_a. It will be noted that e_{ii} is an important factor in rectifier duty and that its effect increases rapidly when the rectifier is controlled by phase retard. Figure 2 shows the effect of phase retard on the arc-back factor. Increase of commutating reactance results in a decrease in rectifier duty. This is not a desirable means for reducing duty as increase in reactance results in

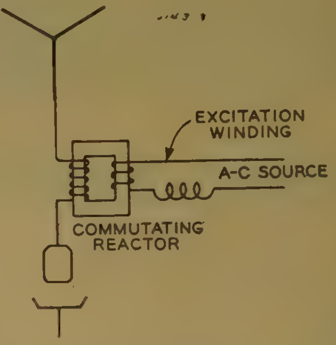

Figure 3. Commutating reactor in anode circuit of rectifier

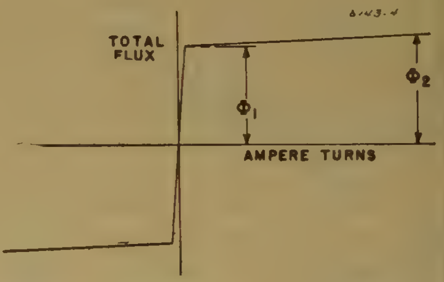

Figure 4. Saturation curve of commutating reactor

increased rectifier regulation and reduced power factor. If the anode current can be "dragged out" or maintained at some low value for a sufficient period just before the end of commutation, a substantial reduction in residual ionization and arc-back probability should result.

Action of Commutating Reactor

A commutating reactor is defined as a reactor used primarily to modify the rate of current transfer between rectifying elements.

A closed core anode reactor having an a-c excitation winding, as shown in Figure

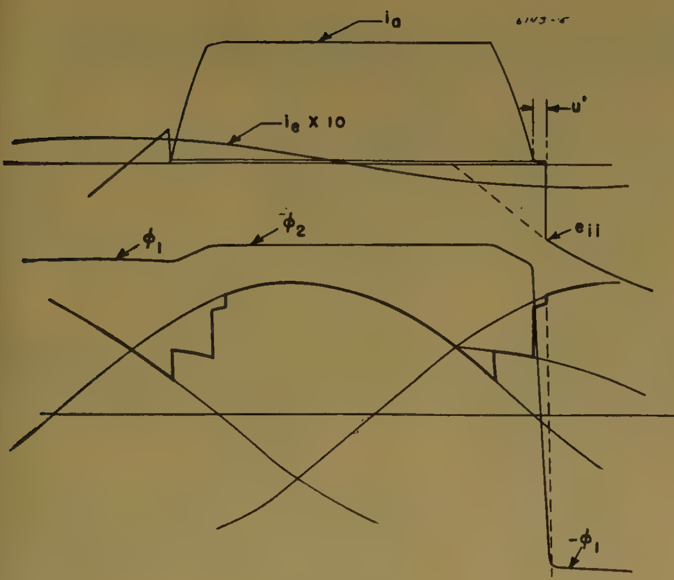

Figure 5. Circuit action of commutating reactor

3, may be used to "drag out" the anode current. Such a reactor preferably would have high permeability below saturation and little flux change beyond saturation, as shown in Figure 4.

Figure 5 shows the circuit action of the reactor. When anode conduction begins, the reactor already is saturated by the excitation winding in the same direction as the anode current, so that the reactor adds relatively little to the commutating reactance. During the anode conducting period, the excitation current i_e reverses.

When commutation to the next anode occurs, it will proceed normally until the anode ampere turns exceed the excitation ampere turns just sufficiently to maintain saturation of the core in the anode current direction. Further decrease in anode current results in desaturation of the reactor. The commutating reactance then will increase, and the anode current will fall slowly at a rate determined by the saturation curve of the reactor and the commutating voltage. This will continue for a time u' until the difference between anode ampere turns and excitation ampere turns is such that the reactor is saturated in the opposite direction, at which time the anode current will fall rapidly the rest of the way to zero.

If the reactor excitation is near the knee of the saturation curve, so that the "drag out" period u' ends at a low value of anode current, then the residual ionization at the end of the conduction period will be reduced. The amount of reduction will depend on the time u'. The value of u' which is needed depends on the arc-back factor and the rate of ion cleanup, which is a function of the geometry of the anode and grid structure of the rectifier.

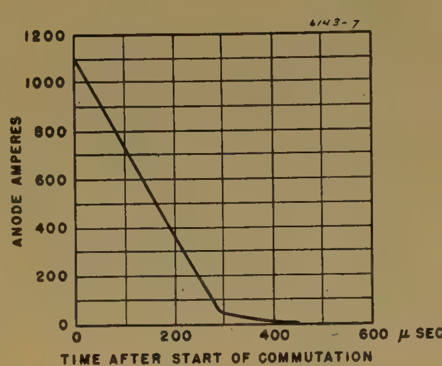

Figure 7. Anode current during commutation with commutating reactor

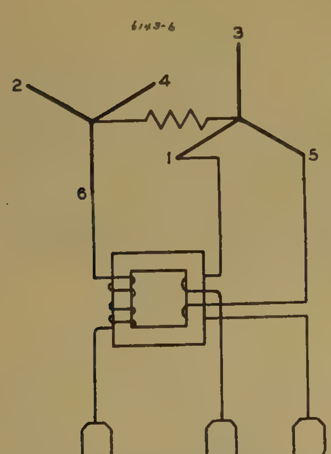

Figure 6. Commutating reactor with excitation by anode currents

Design of Commutating Reactors

The relation $2n\phi_1 \times 10^{-8} = u' \times E_c$ determines the product of anode turns and core section of the reactor. Optimum design requires a minimum ratio of maximum flux to saturating flux in order to reduce the regulation caused by the reactor. The direct voltage loss caused by the reactor is given by

$$E_{xc} = npf(\phi_2 - \phi_1) \times 10^{-8}$$

Since a complete flux reversal occurs in a few electrical degrees, a core design with low eddy loss is needed, otherwise the iron heating will be high and a portion of the drag out period will be consumed in setting up eddy currents in the core structure.

Reactor excitation may be obtained from a separate a-c source, as shown in Figure 3. In this case an impedance is needed in the excitation circuit to limit coupling between the anode circuit and the excitation circuit during the period u', when a high voltage may be induced in the excitation winding. Excitation also may be derived from the currents in other anodes, as shown in Figure 6. If the number of turns in the main winding is so small that this arrangement produces too many excitation ampere turns, suitable current transformers may be used between the anode circuit and the excitation winding.

Operating Results

Figure 7 is typical of the anode current with a commutating reactor in the anode circuit. A record of this current was obtained by placing a small air core reactor in series with the commutating reactor and measuring the amplified voltage across the air core reactor with a cathode ray oscilloscope and camera. A 1,000-cycle sweep was used and was tripped once every 1/60 second so that the record covered 1,000 microseconds. The voltage across the reactor

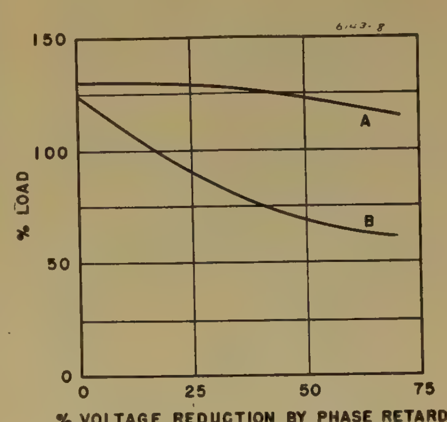

Figure 8. Short time rating of rectifier

A—Commutating reactors
B—No commutating reactors

is proportional to the rate of change of anode current. Since the current at the beginning of commutation was known, the current during commutation could be calculated from the measured rate of change.

Figure 8 indicates the gain resulting from the use of commutating reactors when considerable phase retard is employed. Curve *A* taken with commutating reactors with a drag out time of 350 microseconds at 70 per cent voltage reduction shows that no reduction in rating is needed for phase control. However, a reduction of 40 per cent in rating must be applied in going from 15 to 70 per cent or more voltage reduction without reactors. This is shown in curve *B*. The base load for these curves is the maximum load that can be carried with 15 per cent voltage reduction for a definite time without arc-back.

In this instance it is seen that commutation arc-backs limit the rectifier load both at full and reduced voltages. A minor gain is realized by the use of commutating reactors at full voltage. This becomes a major gain where considerable voltage control is required.

Conclusion

Present day applications of rectifiers call for increasing amounts of voltage control, in many cases to zero voltage. Sustained operation of a rectifier with considerable voltage control usually requires derating of the rectifier. The use of commutating reactors eliminates such derating. They also may prove useful for other cases where high rates of commutation are involved, as at frequencies above 60 cycles.

Appendix. Nomenclature

e_{it} = initial inverse voltage of anode at end of commutation

di_a/dt = final commutation rate of anode current

u' = drag out time (seconds) of anode current caused by action of commutating reactor

n = turns in anode winding of reactor

ϕ_1 = total flux in reactor at knee of saturation curve

ϕ_2 = total flux in reactor at peak anode current

E_c = average commutating voltage during drag out period

i_e = excitation current in commutating reactor

X_c = commutating reactance per phase

p = number of phases in simple rectifier

f = frequency, cycles per second

Investigation of Porcelain Insulators at High Altitudes

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Synopsis: The use of high voltage apparatus in military aircraft made it necessary to obtain flashover data on capacitor bushings at altitudes up to 50,000 feet. Some of the more pertinent data are presented in this paper which gives a-c and d-c flashover voltages for porcelain bushings having spacings of 5/8 inch to 2 7/8 inches from cap to ring. The tests were made at both normal atmospheric and reduced pressures, but no attempt was made to simulate the frosting conditions as encountered in flying. Correlations are shown between bushing spacings and the effect of pressure change. A single correction curve for pressure changes cannot be used, but a family of curves is required for various spacings of insulators and electrostatic field configuration. This paper shows that the practical use of porcelain insulators is limited to altitudes and voltages well below maximum values proposed for aircraft use. It is recommended that solid insulation be used in such a manner as to obtain the benefit of its puncture strength rather than its flashover at the more extreme conditions of altitude and voltage.

UNTIL RECENTLY the low pressure flashover strength of communication apparatus bushings has received little attention. Previous work, such as that reported by Bellaschi and Evans,[1] covers large gaps and insulators as related to power system operation. More fundamental studies of air strength between various gaps at low air density have been presented by Peek[2] and by Schwaiger and Sorensen.[3]

At the beginning of the war, ignition systems were the chief concern and became rapidly more important with increasing altitudes and severity of operating conditions.[4] This problem was solved by proper use of solid insulating materials.

Concurrently, the tremendous increase in air-borne electric apparatus and the subsequent increase in aircraft distribution voltages brought on many practical problems relative to the operation of all sorts of power apparatus, wiring, and auxiliaries. Several authors[5,6] have measured the permissible breakdown and flashover strength of a variety of gaps and insulators under simulated conditions of aircraft operation. Their work, therefore, was concerned with the adverse affects of dust, icing, temperature extremes, and moisture as well as much greater pressure changes than had previously been studied, and for relatively low voltages only.

Toward the end of the war, certain high voltage air-borne apparatus, particularly radar and communication equipment, began to offer new electrical problems unlike any phase of the previous work. The flashover strength of transformer and capacitor bushings, vacuum tube sockets, and similar apparatus became a major problem. After a cursory investigation, it became apparent that size and weight of insulators required was such that it was impractical to insulate apparatus for voltages over about 5,000 volts at stratosphere aircraft operating conditions, which not only include extreme altitudes, but very rapid changes in temperature and altitude with accompanying moisture condensation.

Pressurization of cabins and apparatus compartments has been accepted as the most satisfactory means of reducing the weight of very high voltage apparatus. There remain, however, many marginal conditions where pressurization may be advantageously or necessarily avoided. Choice of the amount of pressurization

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C. V. Fields, a design engineer in the protective devices engineering department and C. L. Cadwell, a laboratory engineer, are both with Westinghouse Electric Corporation, East Pittsburgh, Pa.

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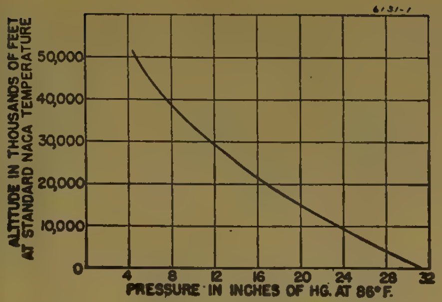

Figure 1. National Advisory Committee for Aeronautics standard altitude corresponding to test chamber pressures at 86 degrees Fahrenheit

or the question of operation in case of pressure failure will depend in part upon apparatus bushing flashover strength. Commercial aircraft will continue to operate at moderate altitudes where pressurization is not warranted. Television and radio equipment will be operated in increasing quantities at moderately high elevations where consideration of the reduction in flashover strength of apparatus bushings caused by reduced pressure will be required.

Previously reported observations of the effects of condensation and frosting on low voltage systems[5] are generally applicable to the high voltage bushings discussed in this paper. That is, unless there is moisture condensation on bushing surfaces, variations in humidity do not affect the breakdown appreciably. Neither is the breakdown affected by frozen moisture. It is to be noted that such factors are of less importance in commercial applications where variations are neither as extreme nor as rapid, and moisture condensation, therefore, less likely to occur. The present work is concerned primarily with the effects of pressure change. All tests were made at room temperature with normal variations in humidity. Omission of a humidity correction seems justified in view of the previous work mentioned.

The National Advisory Committee for Aeronautics standard pressure, temperature, and altitude table generally is accepted for aircraft work. This table gives a standard pressure and a corresponding temperature for each elevation. The curve shown in Figure 1 gives the pressure in inches of mercury (at 86 degrees Fahrenheit) having the same relative air density *RAD* as the standard altitudes shown as ordinates. Thus the measured pressure of 4.7 inches is seen to correspond to a standard elevation of 50,000 feet, which is in turn equal to 3.4 inches of mercury at the standard temperature of −66 degrees.

This transition from room temperature test conditions to high altitudes is permitted by the principle of similitude,[2] which states in effect that for constant spacing the breakdown voltage will be constant if the ratio of P/T is kept constant, where P and T are the absolute pressure and temperature respectively. Since flashover is substantially proportional[7] to relative air density, the following expression may be used to convert our manometer readings taken at the test temperature of 86 degrees to other temperatures and pressures:

$$RAD = \frac{17.95 \times \text{barometric pressure (inches)}}{460 + \text{temperature (degrees Fahrenheit)}} \qquad (1)$$

Test Equipment and Procedure

Most low pressure work previously reported has been conducted with the specimens in a nonmetallic container to reduce field distortion to a minimum. Since it is desirable to test bushings in their normal position mounted on the capacitor, we found it necessary to use a relatively large test chamber. Also, since capacitors usually are surrounded by metal surfaces in practice, it appeared logical to use a metal rather than a glass chamber used by some previous workers testing bushings alone. Capacitor specimens were placed approximately in the center of the tank.

The high voltage lead was brought into the tank through a single hole in the center of the 7/8-inch thick glass cover. The capacitor case was grounded to the tank in bushing-to-ground tests, while the case was insulated and one bushing was grounded for the bushing-to-bushing tests.

The direct voltage was supplied through a 175,000-ohm resistor by a well-filtered rectifier source. A voltmeter was connected directly across the specimen and a dip in the voltmeter reading was taken as the breakdown point. The current was so limited that no apparent damage occurred to the specimen, and a large number of successive readings could be taken without reduction in flashover strength.

The a-c tests were taken on two standard test sets, each with a tertiary voltmeter winding. All points under 50 kv were obtained with a 5-kva 50-kv test set using a hand operated induction regulator. All points above 50-kv were taken with a 50-kva 100-kv test set with a motor driven induction regulator. The wave form conformed to the ASTM Standard D 149.

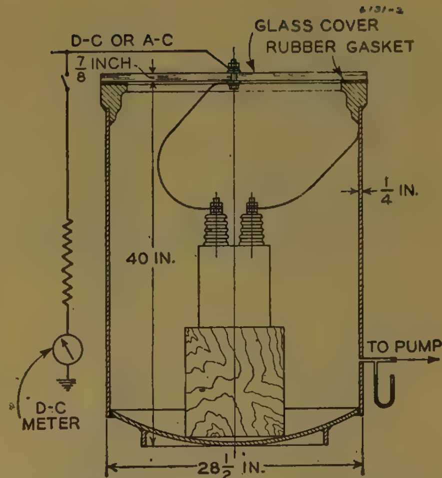

Figure 2. Apparatus for determining capacitor bushing flashover in air at various pressures

Both alternating and direct voltages were raised uniformly at a rate of approximately 10 kv per minute except in the case where the motor driven test set was used, the voltage of which was raised 3 kv per second.

Pressures were measured by a closed tube manometer and a pressure gauge calibrated in inches of mercury absolute. Corresponding altitude values may be obtained by reference to Figure 1 as explained.

Bushing Specimens

These tests were conducted on six typical capacitor bushings shown in Figure 3. All except the gasket sealed bushing *A* were of the "Solder-Seal" metal to porcelain type. They were mounted on conventional capacitor cases with working elements omitted. Leads attached to the top of the bushings were allowed to assume normal positions touching the inside walls of the bushing at random. The leads were clipped off one inch below the bottom of the bushing and were prevented from flashing over inside by filling the capacitor case with Inerteen.*

The bushing size and location and the case dimensions are all commensurate for typical d-c capacitors designed to operate at sea level. Table I gives the major dimensions of the bushings shown in Figure 3 and the covers on which they were mounted.

Summary of Data

The mounting ring on the "Solder-Seal" bushing presents a smooth well-rounded surface, while the cap has rela-

* Chlorinated diphenyl.

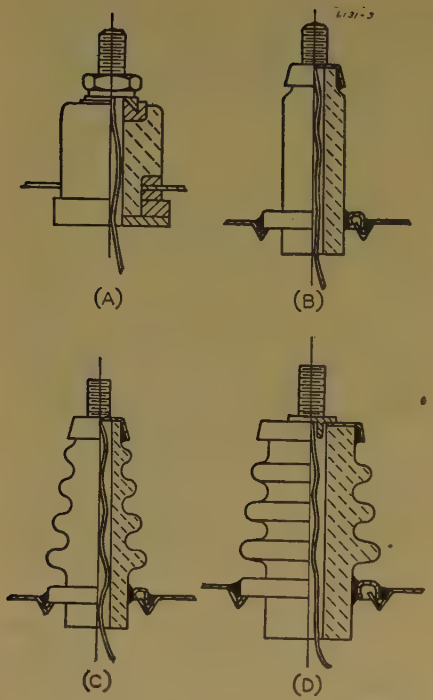

Figure 3. Capacitor bushings tested

Dimensions are shown in Table I

Table I. Major Dimensions of Bushings Shown in Figure 3 and the Covers on Which They Were Mounted

Bushing	Figure	Flashover (Tight-String) Distance in Inches	Diameter (Maximum) in Inches	Striking Distance Between Bushings in Inches	Case Size in Inches		Distance Between Bushing Centers in Inches
					Length	Width	
A	3A	21/32	3/4	11/16	2 1/2	1 3/16	1 1/8
B	3B	1 3/16	3/4	1 11/32	3 3/4	1 3/4	2
C	3B	1 7/8	15/16	1 1/32	3 3/4	1 3/4	2
D	3C	2 1/16	1 1/2	31/32	4 9/16	3 3/4	2
**E*	3D	2 3/16	2 3/4	2 9/32	8	4	4 1/2
				4 21/32	13 1/2	4 1/8	6 3/4
F	3D	2 15/16	2 3/4	2 5/32	8	4	4 1/2
				4 17/32	13 1/2	4 1/8	6 3/4

* Bushing *E* is similar to *F* except that the former has only two corrugations.

tively sharp edges but no burrs. Nevertheless, most streamers seem to form on the lower ring, and breakdown voltages seem relatively unaffected by rounding of the edges on the cap. This suggests that the high field intensity at the base of the bushing was a controlling factor in the breakdown mechanism. Most flashovers were observed to occur between the cap and the mounting ring at 1/16 inch or more from the porcelain and rarely seemed to follow the bushing surface.

The bushings tested differ in form as well as height; however, each is of typical configuration for its height. It seemed logical, therefore, to plot curves as a function of height although it is known that the breakdown is partially dependent on diameter and shape. To have all bushings straight sided and of one diameter would make the test academically correct, but of less practical value. Tight-string distance was measured between cap and mounting ring.

No special efforts were made to avoid ordinary variations in the height of liquid or position of the wire inside the bushings, or other normal factors affecting the breakdown voltages. The relatively large spread in flashover, compared to sphere gaps tested in a uniform field, attests to the effect of these variables. Especially at very low pressures, these normal variations in construction may cause unexpectedly large variations in flashover. The data are sufficiently consistent to warrant drawing the curves shown, but seemed to indicate the desirability of testing capacitors or similar apparatus at actual working pressures.

Bushing to case flashover tests were made first on all six bushings with a-c and both polarities of direct voltage. The data in Figure 4 were obtained by evacuating the chamber, then making a series of tests at four or more pressures, each time admitting air to obtain the pressure desired and taking at least five flashover readings after the pressure had become stable at the new value. All points were plotted and smooth curves were drawn through them. Similar smooth curves were drawn for a-c and positive d-c tests on each bushing. However, only the data for the d-c negative flashover tests are shown since they happen to be more conservative at the lower pressures than the values obtained for the d-c positive flashover tests.

Points corresponding to each of the pressures on the curves of Figures 5 and 6 were taken from the smooth curves drawn through the original test points. These latter curves of flashover versus striking distance, therefore, ignore differences between the various bushings, but show clearly how spacing affects flashover at the various pressures. From these curves it is seen to be increasingly difficult to develop high flashover voltage by increasing the bushing size at greatly reduced pressure. The curve for 4.7 inches of mercury clearly illustrates why about 5,000 volts is a logical maximum voltage for any apparatus bushing operated at or near 50,000 feet. An enormous increase in bushing size would be required to double the flashover voltage.

It is noted that the crest a-c values are somewhat lower than either positive or negative d-c breakdown values at

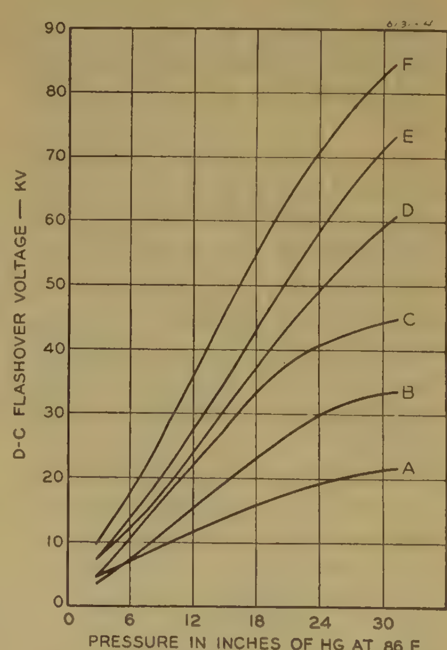

Figure 4. Negative d-c flashover–pressure curves for six bushings tested

Dimensions are shown in Table I

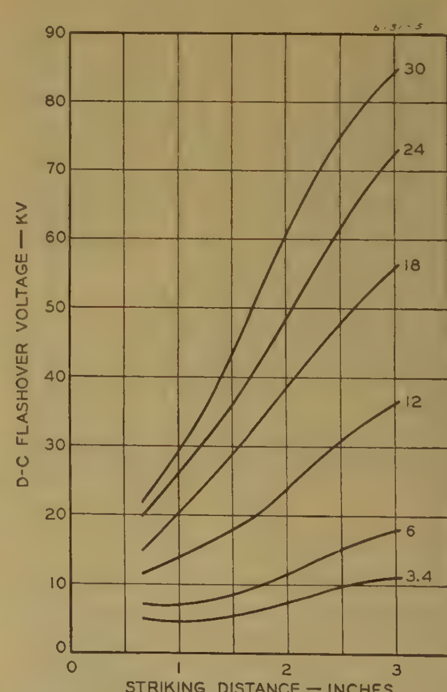

Figure 5. Negative direct voltage required to flash bushings over to case

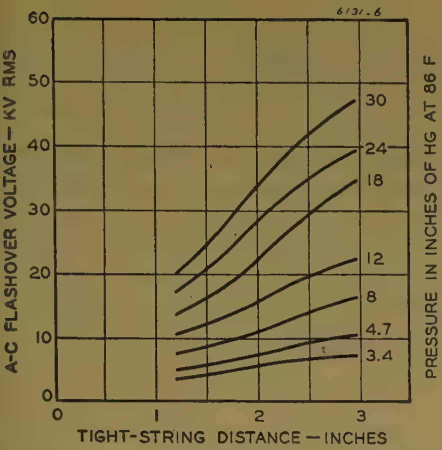

Figure 6. A-c terminal-to-case flashover voltage

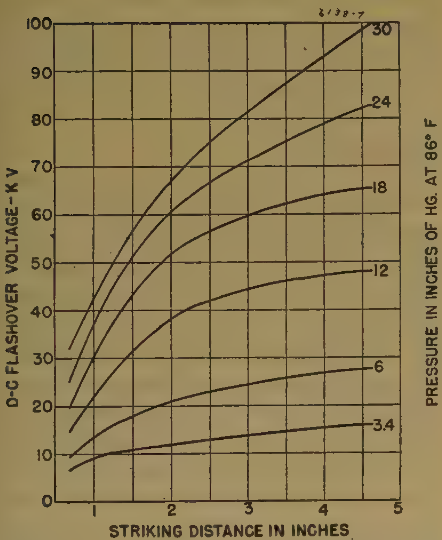

Figure 7. Negative d-c bushing-to-bushing flashover versus spacing

high pressures. This discrepancy is contrary to Berberich's experience in which the peak a-c is higher than d-c values.[5] It is believed that this apparent discrepancy is caused by differences in configuration, both of the field and of the bushings themselves.

A second set of tests was made with the same capacitor bushings mounted as described previously, except that the voltage was applied from bushing to bushing instead of from bushing to case. The flashover values are shown in Figures 7 and 8. Electrostatic fields were assumed to be largely responsible for the considerable rounding of these curves at large spacings.

In order to check the possible relationship between sphere gaps, point gaps, and bushings, the curve in Figure 9 was drawn. From this curve it can be seen that while sphere and point gap flashovers are practically straight lines at sea-level, they flatten out considerably at lower pressures. The bushing to bushing flashover values at 3.4 inches of mercury are intermediate between point and sphere gap values, as might be expected. However, one would not expect the bushing flashover values to fall below those of a point gap. This peculiarity also is believed to be caused by nonuniformity of the field surrounding the capacitor bushings. This explanation is substantiated by a wide variation in flashover voltages of several bushings spaced about the same distance apart, but of different heights.

Ratios of from three to five between sea-level and 45,000-foot altitude flashover voltages found in our work are higher than the 2.7 to 3.7 ratios found by DeLerno[6] in working with smaller gaps.

Table II gives the d-c rated voltage of several capacitors and the flashover voltages taken from the curves for bushings of a size which one would expect to use for high altitude operation. As can be seen, the actual flashover voltages at 45,000 feet are well above the rated values. For comparison, a column is added showing the sea-level test voltage of twice the rated voltage, times 3.7, plus 1,000 volts. It is obvious that the substitution of our maximum ratio of 5 for the 3.7, giving ten times rated voltage plus 1,000 volts, would be nearer the actual sea-level flashover of these particular bushings. However, such a rule would have to be qualified to account for the change in ratio which increases rapidly at large spacings. Furthermore, these sea-level voltages are far higher than could be withstood by the solid insulation used in practice on most apparatus.

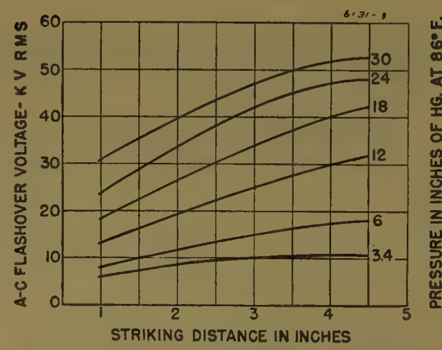

Figure 8. A-c bushing-to-bushing flashover voltage versus spacing

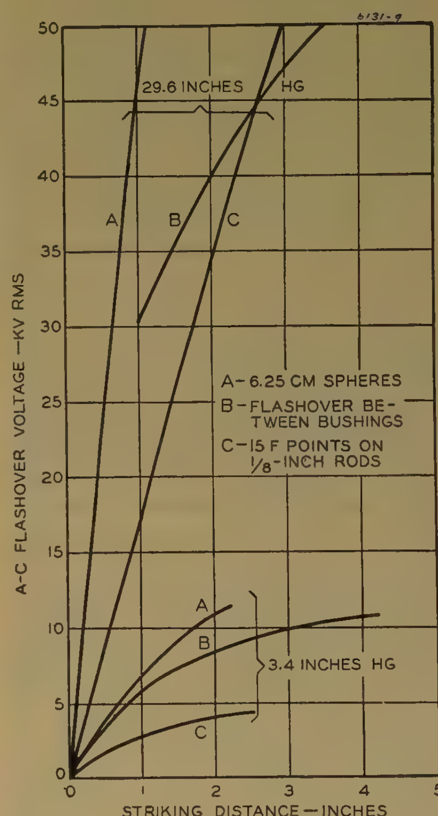

Figure 9. A-c bushing-to-bushing flashover voltages showing relationships to sphere and point gaps with corresponding spacings

Conclusions

1. Our investigations indicate that changes in flashover voltages do not vary in any simple manner with changes in pressure. Ratios of low pressure to high pressure flashover voltages vary considerably with bushing size and configuration, surrounding field, and other factors so that very definite relationships cannot be established.

2. Since the sea-level flashover voltage of most bushings designed for operation at 50,000 feet is far higher than can be withstood by the solid insulation of the apparatus, it appears best to test the completed apparatus in a low pressure chamber to assure proper safety factors at high altitudes. This is particularly true for capacitors because the location of the lead wire and the height of the liquid inside the bushing, and other factors markedly affecting the field change the flashover, especially at very low pressures.

3. The flashover voltage-distance curves for very high altitudes (50,000 feet) are so flat that it is apparent that no reasonable spacing will permit the use of bushings or

Table II. Flashover and Test Voltages for Several High Altitude Capacitor Bushings

Rated Direct Voltage	Bushing Flashover Voltage Distance in Inches	Expected Flashover Voltage at 45,000 Feet*	Actual Flashover Voltage at Sea-Level	Twice Rated Voltage Times 3.7 Plus 1,000 Volts
1,500	13/16	7,000	25,000	12,000
3,000	1 3/16	7,500	33,000	23,200
5,000	1 13/16	10,500	57,500	38,000

* Computed for NACA standard altitude from curves taken at 86 degrees Fahrenheit.

Diesel-Electric Synchronous Motor Ship Drives—Propulsion Control Problems

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Synopsis: Although the many obvious advantages of ship propulsion drives with synchronous motors and generators, and with power supplied by Diesel engines, have been realized for some time, no actual installations were made in this country until a few years ago when a trial installation was completed on a 12,000 horsepower twin-screw vessel. Experience gained from operation of this vessel has established that this type of drive in many respects is superior to other drives, and indications are that it may find extensive use on future vessels. The paper discusses control problems associated with the drive and describes methods of control and types of control equipment which have been found best suited for the drive.

DURING the last 25 years, a rapidly increasing number of naval and merchant vessels has been equipped with electric propulsion equipment, until by now more than 2,000 American vessels are electrically propelled. Primary power for the propulsion motors is supplied by Diesel engines or steam turbines coupled to electric generators. Up to the present time, with very few exceptions, all Diesel engine powered ships have d-c generators and motors, while synchronous generators and motors are used for steam-turbine drives.

The third possibility of using synchronous machines and Diesel engine prime movers for propulsion has been under considerable discussion for years. This type of drive has many obvious merits, but anticipated operating difficulties, especially during maneuvering, delayed actual installations on American vessels for several years. Finally in 1940, with the encouragement of various manufacturing concerns, plans were prepared and an order placed for a trial installation on a twin-screw vessel to demonstrate the practicability of the drive.

The vessel, which was completed in 1942, has two 5,900-horsepower 2,400-volt 3-phase synchronous propulsion motors. Power is supplied by eight 1,600-horsepower Diesel engines direct connected to 1,150-kva generators. The experience gained during sea trials and, through actual operation of the vessel for three years has proved conclusively that this type of drive not only is practical but in many respects is superior to similar drives using d-c equipment. There also is a good possibility that it may offer strong competition to turbine-electric drives for many applications.

Among the advantages of the drive, the following can be mentioned:

1. Synchronous machines weigh less and cost less than d-c machines of the same rating and their efficiency is higher.

2. Synchronous machines are more reliable and easier to maintain than d-c machines.

3. Machines with higher voltage can be used, thereby causing a reduction in cables and bus bars and permitting installation of units with higher ratings than is practical with d-c machines.

4. High speed Diesel engines with greatly increased ratings, now under development, will permit installation of high power Diesel-electric drives on vessels ordinarily powered by steam turbines, thereby dispensing with the bulky and highly vulnerable boiler plants on these vessels. The use of two or more Diesel-electric generating units instead of one larger turbine-electric unit will give additional flexibility and reliability to the propulsion drive.

In view of the above, it is commonly believed that Diesel-electric synchronous motor propulsion drives will be used extensively on future electrically propelled vessels. A brief discussion of problems, type of equipment, and method of control applying to the drives, therefore, should be of interest to a great many people who in some way are connected with the building or operation of ships.

Diesel Engines

Diesel engines for propulsion service, through constant improvements, have attained the high degree of dependability, efficiency, and durability required for this service. The output and speed of the engines have been raised steadily, and engines with a continuous rating of 2,000 horsepower at 800 rpm are now available. Engines of considerably higher rating and speed are under development and should be available in the near future. The engines operate satisfactorily at speeds as low as 30 per cent of rated speed. They develop rated torque down to 50 to 60 per cent speed, but at lower speeds the torque will be reduced gradually to as low as 60 per cent of rated torque at 30 per cent speed.

The engine speed is controlled by governors of the hydraulic type. As a protection against overloading, the governors will automatically limit the amount of fuel which is admitted to the engine per stroke. This amount decreases at low speeds so that the maximum torque, at any speed, will not exceed a safe value for continuous operation. If the torque is exceeded the engines will slow down and unload themselves.

The speed of the engines is adjusted remotely from the propulsion control board. A master speed control transmitter, coupled to the speed lever, serves

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air insulation appreciably above 5,000 volts. Connectors taking advantage of solid insulation puncture strength, similar to those used in aircraft ignition systems, should be developed if higher voltages are encountered at extreme altitudes.

4. The curves shown represent the conservative averages of a great many tests made under practical conditions, and it is believed that they will serve as a guide in the selection of porcelain bushings falling within the scope of our work.

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to position receivers on the engine governors. Pneumatic or electric positioning devices are available for the speed control. Since the engines must operate in parallel at the same speed, means must be provided to equalize the engine loading. This may be performed automatically or by means of manual vernier speed control transmitters on the control board.

Starting and reversal of the propulsion motor are performed with the engines running at minimum speed. For satisfactory performance the engines must develop the highest possible torque during this period. The torque limitation, imposed on the engines by the governor action, therefore, is removed by energizing of magnetic solenoids provided on the engine governors for this purpose. This will raise the engine torque to a maximum value on the order of 85 per cent of rated torque, which can be delivered for short periods without causing damage.

Motor Starting Torque

During maneuvering periods, when the propulsion motor is being started or reversed, the motor operates as a squirrel cage induction motor with the de-energized field connected in series with a starting resistor. To develop the necessary starting and synchronizing torque, the motor must be provided with heavy damper windings and the associated generators must be overexcited with 200 to 300 per cent normal field current. The motor torque is a combination of two individual components. One component is produced by the current induced in the damper winding while the other component is produced by the current induced in the field coils. The value of the latter component is a function of the value of the starting resistance.

Typical curves, representing the two torque components and the resultant torque as a function of the resistor value, are shown in Figure 1 for 30 per cent generator speed. It will be noted that the peak value of the field torque components *B*1, *B*2, and *B*3 is practically independent of the resistor value but occurs at different motor slips. The starting resistor value should be selected so that a maximum resultant motor torque will be developed at the highest motor slip at which successful synchronizing is possible. The resultant torque curve *C*2 fulfills this requirement. Under certain conditions it may be necessary to bring the motor up to a maximum speed with one resistor value, and then increase the speed further by shorting out part of the resistor before synchronizing is possible. The peak value of the combined torque occurs at 35 to 50 per cent motor slip and should not be less than 50 to 60 per cent of full load torque when all generators are in service and overexcited.

Torque Margin

Torque margin for an electric propulsion drive is defined as the difference between the torque required by the propeller for normal full power running and the torque at which the motor pulls out of step, expressed as a percentage of the propeller torque.

For Diesel-electric drives, a torque margin of 10 to 15 per cent at normal generator and motor excitation is sufficient. This low torque margin is permissible because of the torque limiting feature of modern engine governors. When operating the vessel in rough weather or during periods of acceleration the propeller torque requirement frequently exceeds the pull-out torque. Since the engines cannot develop a torque of this value, however, the result is only a momentary reduction in engine and motor speed and a corresponding reduction in the propeller torque requirement, but the motor will not pull out of step. The transient increase in propeller torque builds up at a relatively slow rate, and the inertia of engines and generators is so low that the engine speed reduces fast enough to prevent the actual occurrence of torques higher than the pull-out value.

This is in contrast to conditions encountered on steam turbine drives. The torque developed by a turbine cannot conveniently be limited by governor action, and may easily exceed rated torque. Furthermore, the turbine inertia is relatively high and its speed cannot be reduced fast enough to prevent occurrence of transient torques, exceeding the pull-out torque, even though the turbine torque was limited. For this reason, the torque margin built into motors and generators for turbine-electric drives must be increased, causing an increase in weight and dimensions, or provisions must be made to increase automatically the pull-out torque whenever required. This is accomplished by a momentary increase in motor and generator excitation.

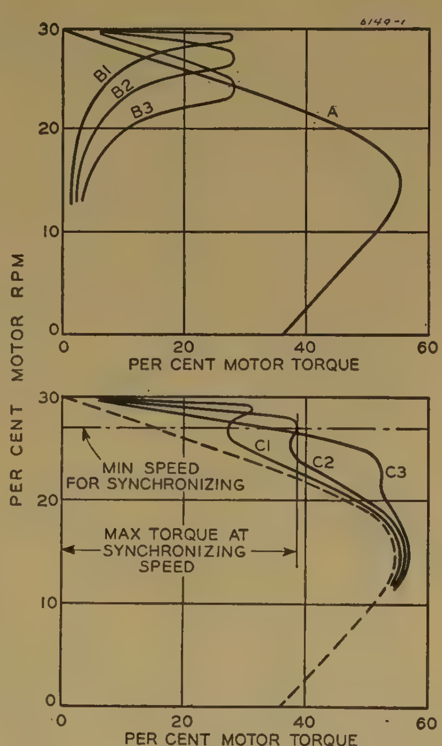


Figure 1. Curves showing the starting torque of a synchronous motor as a function of motor speed for different values of starting resistance

A—Torque developed by damper winding
B—Torque developed by motor field winding
*B*1—Low starting resistance
*B*2—Medium starting resistance
*B*3—High starting resistance
C—Total motor torque
*C*1—Low starting resistance
*C*2—Medium starting resistance
*C*3—High starting resistance

Dynamic Braking

Typical curves for propeller torque requirements during reversal of a vessel with four propulsion generators are shown in Figure 2. It will be noted that the peak propeller torque, which occurs before the motor stops, actually is higher than the torque required to hold the motor at standstill. This peak torque is of particular interest because it occurs at a point where the available motor torque is low and, in most instances, is not sufficient to overcome the peak torque.

To illustrate this condition, a series of curves representing the available motor torque for two, three, and four generators, operating at 30 per cent speed, have been shown for comparison. Examination of the curves discloses that the propeller cannot be stopped by motor torque alone, even with four generators in service, and that the condition is equally adverse for 2-generator operation, although the propeller torque is reduced, as a result of the lower ship speed in this case.

To stop the propeller preparatory to a reversal, two methods are available:

1. Permit the vessel to slow down without power on the motor until the propeller torque is reduced below the corresponding motor torque at all speed points, and then

apply reverse power. The time required for the slow-down, however, is excessive, especially for large vessels, and this method is not acceptable.

2. Stop the motor by dynamic braking. By this means the motor can be brought to practically standstill and reverse power applied in a few seconds. Since the additional equipment, required for dynamic braking, is relatively inexpensive, this method should be incorporated on all drives of this type as a means for reducing the time required for motor reversal.

Dynamic braking of the motor is produced by connecting a 3-phase resistor to the motor terminals and energizing the motor field with normal field current. The effect of the value of the braking resistance on the braking torque is illustrated in Figure 3, where curves are shown for braking torque as a function of motor speed for three different resistance values. It will be noted that the value of the peak braking torque is practically independent of the resistance value. However, the torque peak occurs at a higher motor speed when the resistance value is increased.

The propeller torque curve for maximum ship speed also is shown to illustrate the relation between this torque and the braking torque. The points of intersection of the curves give the minimum value to which the motor speed will be reduced for different values of braking resistance. The speed decreases with decreasing resistance value. For very low resistance values, however, the braking torque is not sufficient to overcome the propeller torque at higher motor speeds, as illustrated by curve $A1$, and the motor speed will not be reduced below speed C. This must be kept in mind when designing the starting resistor, and a check should be made to ascertain that the braking torque is higher than the motor torque at all speeds, as illustrated by curve $A2$.

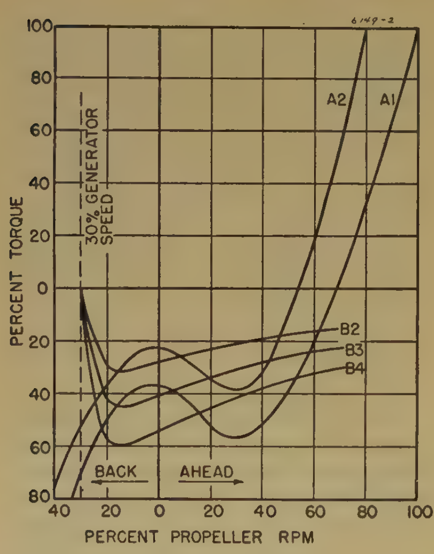


Figure 2. Curves showing motor and propeller torque as a function of motor speed during reversals; vessel moving at full speed ahead

$A1$—Propeller torque, four generators in service
$A2$—Propeller torque, two generators in service
B—Torque developed by propulsion motor
$B2$—Two generators in service
$B3$—Three generators in service
$B4$—Four generators in service

General Control Scheme

A complete propulsion plant consists of two or more Diesel engines directly connected to synchronous generators, a synchronous propulsion motor, an excitation supply, and the propulsion control equipment.

The method of control is very simple and should not vary greatly for different installations. The general description given in this paper gives a fairly complete picture of the type of equipment, control scheme, and operating procedure for most installations. A simplified schematic diagram of a typical control scheme is shown in Figure 4 for a synchronous drive with three generators.

Any combination of the generators can be connected to a common generator bus by means of three high voltage contactor groups. The contactors for generator number one are numbered $1G1$, $1G2$, and $1G3$. The motor can be connected to the same bus by means of reversing contactors $R1$ to $R4$, of which $R2$ and $R3$ close for "ahead" operation and $R1$ and $R4$ close for "astern" or "back" operation. Quick stopping of the motor is made possible by connecting a dynamic braking resistor to the motor terminals by the closing of contactors $R5$ to $R7$.

The excitation for the propulsion machinery is supplied by a motor-generator set with separate exciters for motor and generator excitation. The exciter fields are energized from a constant voltage control bus through a magnetic field contactor F, which remains closed for normal operation. The generator and motor exciter fields are energized by the closing of starting switches $S4$ and $S6$. The generator and motor fields are connected to the associated exciter busses by means of field switches as shown. When a field is de-energized, it is connected to a discharge or starting resistor to prevent high inductive voltage peaks.

The cooling of the field coils becomes less effective when the speed of the propulsion motors and generators is reduced. The exciter voltage, therefore, must be decreased for low speed operation, by means of rheostats in the exciter field circuits, to prevent overheating of the coils. The rheostats are coupled to the speed control lever, and so designed that the exciter voltage will be reduced to approximately 75 per cent of normal when the speed lever is moved from the "fast" to the "slow" speed position. The rheostats are shorted by means of starter switches $S5$ and $S7$ during the starting period when maximum excitation is needed to obtain the required starting and synchronizing torques. The generator exciter voltage must be raised to 200 to 300 per cent of normal during this period. A resistor is connected in series with the generator exciter rheostat to reduce the exciter to voltage to normal when switch $S5$ opens in the "run" position.

For the sake of simplicity it is assumed that all contactors and switches, shown in the diagram, are of the cam operated type arranged in groups and controlled manually by means of operating levers in a control stand at the front of the control board. A diagram of the control stand is shown in Figure 5.

Incorrect operation of the starting levers is prevented by means of mechanical interlocking as indicated in the lever diagram. The purpose of the interlocking is as follows:

1. The motor speed lever and the starter lever are interlocked to prevent starting of the motor unless the engine speed is a minimum. It also is necessary to reduce the speed to a minimum before stopping the motor.

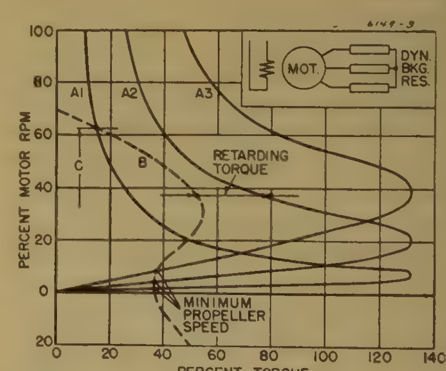
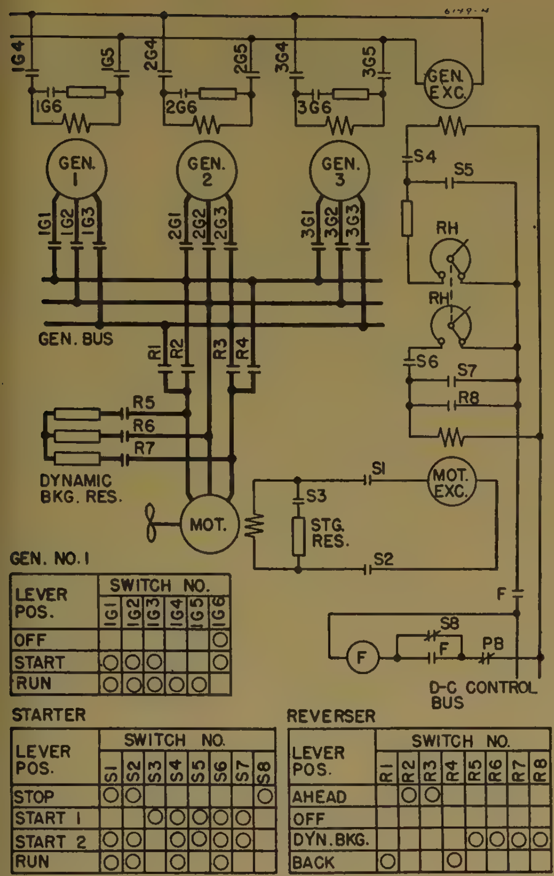


Figure 3. Curves showing dynamic braking torque as a function of motor speed for different values of braking resistance

A—Dynamic braking torque
$A1$—Low dynamic braking resistance
$A2$—Medium dynamic braking resistance
$A3$—High dynamic braking resistance
B—Propeller torque, vessel moving at full speed ahead



GEN. NO. 1

LEVER POS.	SWITCH NO. 1G1	1G2	1G3	1G4	1G5	1G6
OFF						O
START	O	O	O			O
RUN	O	O	O	O	O	

STARTER

LEVER POS.	SWITCH NO. S1	S2	S3	S4	S5	S6	S7	S8
STOP	O	O						O
START 1			O	O	O	O	O	
START 2	O	O		O	O	O	O	
RUN	O	O		O		O		

REVERSER

LEVER POS.	SWITCH NO. R1	R2	R3	R4	R5	R6	R7	R8
AHEAD		O	O					
OFF								
DYN. BKG.					O	O	O	O
BACK	O			O				

Figure 4. Simplified schematic control diagram for synchronous motor propulsion drive

2. The starter lever and the reverser lever are interlocked to prevent moving the starter lever from the "stop" position, unless the reverser lever is in the "ahead" or the "back" position. Inversely, the reverser lever cannot be moved unless the starter lever is in the "stop" position. Since, in this case, the speed lever must be in the slow speed position, it follows that the reversing contactors cannot be operated unless the generator speed is a minimum and the generator field is de-energized. The operating duty of the contactors, therefore, is considerably lightened.

A speed control transmitter is connected to the speed lever for the purpose of controlling the speed of the engines and, consequently, of the motor over the speed range. Exciter rheostats, required for reduction of generator and motor excitation at reduced speeds, also are coupled to the lever.

Motor Starting

The generators, which will be used for propulsion, are connected to the generator bus by placing the generator levers in the "run" position while the engines are running at idling speed. The generator field switches (*1G4* and *1G5* for generator number one) close and connect the fields to the exciter, which, however, is de-energized since the starter lever is in the "stop" position and switch *S4* is open. The generators, therefore, will not be synchronized but will run at approximately equal speeds determined by the master speed control transmitter.

As the next step, the reverser lever is moved to the "ahead" position, thereby closing contactors *R2* and *R3* and connecting the motor armature to the generator bus. Although motor field switches *S1* and *S2* are closed, the motor field remains de-energized since exciter field switches *S6*, *S7*, and *R8* are open.

Actual starting of the motor takes place when the starting lever is moved to the "start 1" position. Motor field switches *S1* and *S2* open and the starting resistor is connected in the field circuit by the closing of switch *S3*. Switch *S7* closes and applies maximum excitation to the motor exciter in preparation for the motor synchronizing. Simultaneously, the generator fields become overexcited since switches *S4* and *S5* close. Although there may be a considerable difference in the speed of the Diesel engines, the synchronizing torque resulting from the overexcitation is sufficient to pull the generators into step and cause synchronization. Following this, the motor starts up as a squirrel-cage induction motor and accelerates gradually to near synchronous speed.

When the motor approaches synchronous speed, the main motor current drops sharply from an almost constant high value during the accelerating period and thereby furnishes an indication that the motor may be synchronized. At this point, the starter lever is moved to the "start 2" position. Switches *S1* and *S2* close and connect the motor field to the previously excited motor exciter. Switch *S3* opens and interrupts the circuit to the starting resistor. As a result the motor becomes energized and begins operating as a synchronous motor. After a few seconds hesitation the starter lever is moved to the "run" position. Switches *S5* and *S7* open and reduce motor and generator excitation to normal. The motor speed now may be adjusted to any desired value by means of the speed lever which controls the engine speed control master transmitter.

Discussion of Motor Starting Sequence

The starting sequence, outlined above is unorthodox and probably will be questioned. When first confronted with the problem it would seem more logical first to synchronize and overexcite the generators, connected to the generator bus, and then to start the motor by connecting it to the energized bus. This method is not satisfactory, however, for reasons as outlined as follows.

Before starting the motor, the engines run unloaded at a speed only slightly higher than their minimum stable speed, and the amount of fuel oil, admitted to the engines by the governors, is very low. When connecting the motor to the overexcited generator bus, there will be a se-

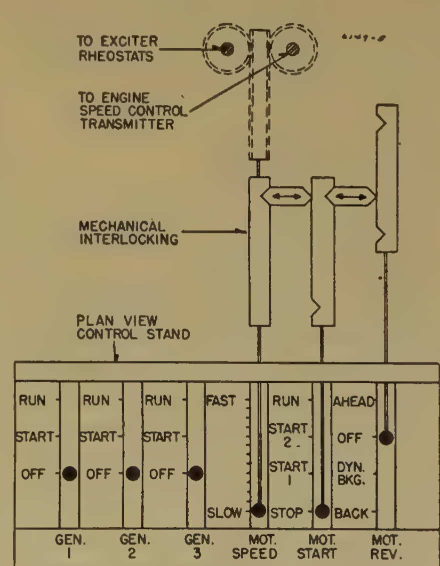


Figure 5. Diagram of lever operating stand with mechanical interlocking

vere and practically instantaneous increase in the generator loading and a corresponding rapid reduction in engine speed. The governors, consequently, will function to increase the amount of fuel oil and the engine torque, but this action is not instantaneous. During the time required to increase the engine torque sufficiently to equal the generator torque, the engine speed frequently will drop below the stable speed and stalling results.

However, when the generators are energized after the motor is connected to the generator bus, the loading increases gradually because of the time constants of exciter and generator fields. The response of the governors is sufficiently fast to follow the load increase, and there is no appreciable reduction in engine speed, and stalling of the engines is prevented.

Motor Reversal

To reverse the propulsion motor when the vessel is making considerable headway, generators and motor first must be de-energized by moving the speed lever to the "slow" position and the starter lever to the "stop" position. The motor now will idle at approximately 70 per cent of normal speed, but is brought quickly to practically standstill by placing the reverser lever in the "dynamic braking" position. Contactors *R*5, *R*6, and *R*7 close and connect the braking resistor to the motor terminals. At the same time, switch *R*8 closes and applies maximum excitation to the motor field.

As soon as the motor stops, the reverser lever is moved to the "back" position and the starter lever is moved successively to the "start 1," the "start 2," and finally to the run position. The motor starts up in the same sequence as described for normal starting except that the phase rotation is reversed.

Control Problems During Reversal

When making a reversal, the speed of the engines and generators must be reduced to a minimum, since the torque requirement of the motor at synchronizing speed otherwise would become excessive. This presents a serious problem because the engine torque is reduced at low speeds, and also because the engine operates so close to the minimum stable speed that a momentary drop in speed may cause stalling.

To improve conditions, the torque limitation is removed during maneuvering by energizing solenoids on the governors. This will raise the engine torque to 85 per cent or more of normal which can be maintained for short periods. The torque requirement, however, exceeds even the increased engine torque during full power reversals (crash stop) and causes stalling unless special precautions are taken. The condition becomes especially critical where a reversal is attempted with a reduced number of engines. Even normal starting from standstill then may cause stalling.

To obtain a clear picture of the problems involved, curves representing propeller, motor, and engine torques as a function of motor speed are shown in Figure 6. Curve *A*1 represents the propeller torque when the vessel is moving ahead at full power speed. *A*2 and *A*3 are similar curves for reduced ship speeds. Curve *B*1 represents the torque developed by the motor at 250 per cent generator excitation and 30 per cent engine speed. *B*2 and *B*3 are similar curves for reduced generator excitation. The torque required by the engines to produce the corresponding motor torque is represented by curves *C*1, *C*2, and *C*3. It should be noted that the motor torque, as shown, is a function only of the electrical characteristics of motor and generators and has no relation to the propeller torque requirements. The difference between motor and propeller torque is the torque actually available for acceleration. The engine torque is a combination of the motor torque and the considerable electric and mechanical motor and generator losses.

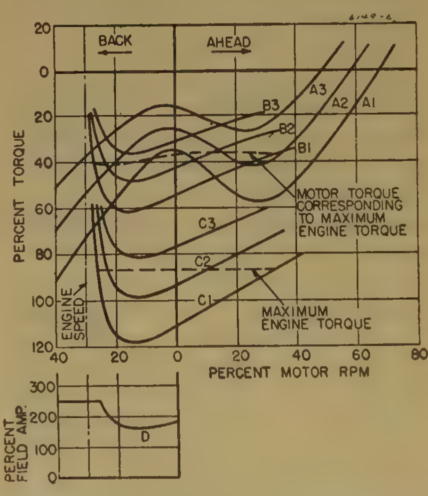


Figure 6. Curves showing propeller, motor, and engine torque as a function of motor speed during reversals

A—Propeller torque
*A*1—100 per cent ship speed
*A*2— 85 per cent ship speed
*A*3— 70 per cent ship speed
B—Propulsion motor torque
*B*1—250 per cent generator excitation
*B*2—200 per cent generator excitation
*B*3—150 per cent generator excitation
C—Engine torques corresponding to above motor torques
D—Generator field current required to maintain maximum engine torque

An examination of the curves discloses that with 250 per cent generator excitation the torque requirement exceeds the maximum available engine torque and the engines will stall. To remedy this condition, the torque requirements must be reduced by reduction of the generator excitation to match the available 85 per cent engine torque. The motor torque available for acceleration of the propeller, of course, is reduced correspondingly, as illustrated.

When reverse power is applied to the motor, after it has been brought to nearly standstill by means of dynamic braking, the reduced torque is sufficient only to raise the reverse motor speed to less than ten per cent where the curve for available motor torque intersects curve *A*1. From this point on, the motor speed increases gradually as the ship speed is reduced, and reaches close to synchronizing speed at 70 per cent ship speed. The generator field current required to maintain 85 per cent engine torque in the accelerating period is represented by curve *D*.

The vessel slows down surprisingly fast as a result of the combined braking effect of hull friction and the reverse propeller thrust, and the time required to accelerate the motor to synchronizing speed is not excessive. The head reach of a vessel with this type of drive compares favorably with the value obtained for similar vessels with d-c drives.

Automatic Torque Control

Satisfactory operation during maneuvering to prevent engine stalling requires automatic control of the generator field current. The current must be maintained at the highest value possible without causing reduction of the engine speed below a permissible minimum value.

A suitable control to meet this requirement is illustrated in the schematic diagram, Figure 7. A small permanent magnet d-c pilot generator *PG* is driven by a 3-phase induction motor, connected to the generator bus through a step-down transformer. The pilot generator voltage is a measure of the main generator frequency and, consequently, of the engine speed. Also coupled to the motor is a regulating exciter *RE* provided with a self-exciting field and a separate control field. A resistor in the self-exciting field circuit is adjusted so that the generator voltage is zero when the control field current is zero, but

will build up to a high value when a very low current is passed through the control field.

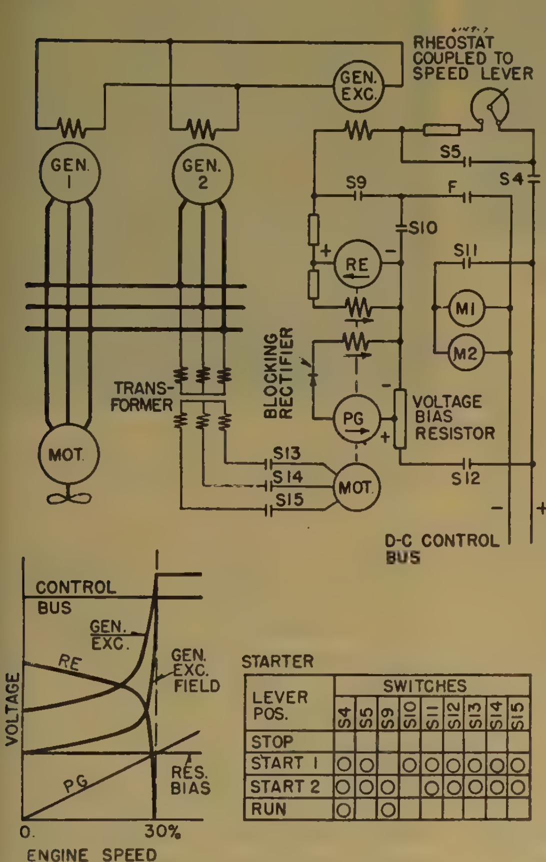
The pilot generator is connected in series opposition to a bias voltage across part of a resistor energized from the control bus, and the voltage difference is applied to the control field. The bias voltage is adjusted to equal the pilot generator voltage when the engine speed is 30 per cent of normal. At higher engine speed, the generator voltage exceeds the bias voltage, but current flow through the control field is prevented by a rectifier block. If the engine speed drops as the result of excessive torque requirement, the bias voltage exceeds the generator voltage and a current will flow through the control field in a direction as indicated. The resulting exciter voltage bucks the control bus voltage in the generator exciter field circuit and reduces the field current. This causes a corresponding reduction in main generator excitation and engine torque until stability is regained. The functioning of the torque regulating equipment is illustrated in Figure 7 by a series of curves showing the relative values of the excitation and control voltages as the engine speed decreases.

Torque control is required only until the motor is synchronized, and the regulating equipment is disconnected when the starter lever is moved to the "run" position by the opening of switches $S10$ to $S15$. The de-energized rototrol is bypassed by switch $S9$. The governor solenoids $M1$ and $M2$, which serve to increase the engine torque during the maneuvering period, are de-energized and normal governor function restored by the opening of switch $S11$ when the starter lever is moved to the "run" position.

Paralleling of Generators

When, during normal operation with one or more generators in service, it becomes necessary to increase the power output by connecting an additional generator to the bus, the first step is to start the associated engine and adjust its speed to approximately the same speed as the engines in service by connecting the governor actuator to the master speed control transmitter. The next step is to move the generator lever to the "start" position, thereby connecting the generator armature to the bus, while the de-energized generator field is connected to a field resistor (see Figure 4). The generator as a consequence, will run as a squirrel cage induction motor and will be forced into near synchronism with the other generators. When, as the third step, the generator lever is moved to the "run" position, the generator becomes energized and pulls into step.

To make possible this simple method of synchronizing, the same consideration must be given to the design of damper winding and field resistor for the generators as for the propulsion motor. With correct design, the synchronizing is very positive and it actually is possible, as has been demonstrated, to start an engine from standstill. If the engine starting lever previously has been placed in the "run" position, the engine will start automatically and begin to deliver power simply by advancing the generator lever to the "start" and then to the "run" position.



LEVER POS.	S4	S5	S9	S10	S11	S12	S13	S14	S15
STOP									
START 1	O	O		O	O	O	O	O	O
START 2	O	O	O		O	O	O	O	O
RUN	O		O						

Figure 7. Schematic diagram of automatic engine torque regulation

RE—Regulating exciter
PG—Permanent field pilot generator
M1, *M2*—Engine governor solenoids

Operation at Subsynchronous Speeds

The minimum propeller speed possible with synchronous motor operation is limited by the minimum stable engine speed to approximately 30 per cent of normal speed. On a 20-knot vessel, the minimum steady speed will be in excess of six knots. This speed is somewhat high, and it is desirable to operate at lower speed on some classes of vessels.

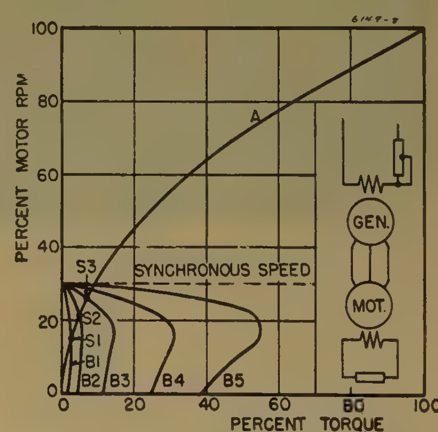


Figure 8. Operation of propulsion motor at subsynchronous speeds

Curves *B1* to *B5* show motor torque as a function of speed for different values of generator excitation. *B1* is for minimum excitation, *B5* is for overexcitation, *A* is propeller torque curve for steady running condition

This is accomplished by operating the motor as a squirrel cage induction motor at subsynchronous speed. The motor speed is controlled by adjustment of the generator excitation below its normal value, while the de-energized motor field is connected to the starting resistor.

Motor torque curves for several values

of generator excitation are shown in Figure 8. The intersections between these curves and the propeller torque curve give the values of the corresponding motor speeds for steady running. To prevent overheating of motor and generators as a result of high magnetizing and damper winding losses, the generator excitation must be reduced considerably below normal value for continuous operation. Operation at speeds between the minimum synchronous speed and 70 per cent of this value, therefore, is not possible. One speed point at approximately 50 per cent of minimum synchronous speed or 15 per cent of maximum motor speed should be sufficient. To permit operation at this speed, the starter lever has an additional position between "stop" and "start 1" positions, in which a field resistor is inserted in the generator exciter field circuit. For satisfactory operation at least 50 per cent of the generators must be in service.

Protection

The power requirement of a propulsion motor is a function of the propeller speed for any given installation with minor changes caused by the condition of the hull. The motors are designed to supply the necessary torque for the maximum propeller speed and, consequently, cannot be overloaded as long as this speed is not exceeded. The Diesel engines are designed to supply the rated generator output within a narrow margin, thereby precluding overloading of the generators.

Serious overloading of an electric propulsion plant, therefore, is not possible, and, since the equipment is under practically constant supervision when the vessel is under way, normal overload protection is not provided. It is only necessary to protect the main circuits against open circuits and short circuits which occur more frequently than on land installation because of the severe operating conditions.

Phase balance relays in each generator circuit are provided as a protection against this type of fault. The relays trip on a phase unbalance caused by short circuits or open circuits, but are insensitive to the high currents in the circuits during maneuvering. The relay contacts are connected in the coil circuit of a magnetic field contactor and will open and de-energize motor and generators when a fault occurs. The main contactors, therefore, are not required to open short-circuit currents. The contact *PB* of a phase balance relay is shown in diagram, Figure 4. To reset the field contactor after the fault has been cleared, the starter lever must be moved momentarily to the "stop" position when switch *S*8 closes.

Figure 9. Typical propulsion control board mounted on engine builder's test floor

Design of Control Boards

The control boards are of the metal enclosed dead front type and should have separate compartments for high and low voltage apparatus if space permits. In accordance with standard marine practice, the main and field contactors are usually of the manually operated type for the sake of reliability. Pneumatically or magnetically operated contactors, however, may be more suitable in many instances, especially on large installations since this permits mounting of switch units in convenient locations remote from the main control station. More positive operation is also possible than with manual switching.

Manual contactors usually are operated by cam shafts, connected mechanically to the operating levers. On installations with pneumatically or magnetically operated contactors, master switches are provided for remote control of the contactors. It may be found desirable to retain the lever operating stand also for this type of control and connect the master switches to the levers. This also simplifies the problem of providing the necessary mechanical interlocking. A control board of this type is shown mounted on the engine builder's test floor in Figure 9.

Figure 10. A 2,500-volt 1,500-ampere cam-operated trip-free propulsion contactor

The high voltage main contactors should be of the air break type. They are required to interrupt normal generator and motor currents but not short-circuit currents. When the contactors are of the manual type, it may be necessary to provide trip-free features on the generator contactors to permit remote tripping from the engine gauge boards of associated contactors in case of engine failure. A typical 2,500-volt 1,500-ampere cam-operated trip-free type contactor is shown in Figure 10.

Rectifier Capacity

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THE LOAD CAPACITY of mercury arc rectifiers must be known in order to utilize them to fullest extent and best advantage. Knowledge of the factors affecting rectifier capacity and the relations between them has increased so that it is now possible to specify the rectifier capacity in definite terms.

Mercury arc rectifier tubes (or tanks) are manufactured only in a limited number of types and sizes. A large variety of rectifier units of different ratings and characteristics may be made up from combinations of one or more similar tubes. Information regarding the capacity and characteristics of each tube (or tank) is essential to the design and application of such mercury arc power converters. It is the purpose of this paper to describe the load-time and volt-ampere characteristics of typical mercury arc tubes of the ignitron type and discuss the relations between these characteristics and the factors affecting them. The control and other characteristics of mercury arc tubes will be treated in a companion paper.[1]

The load carrying ability of a mercury pool tube may be limited by arc-back, voltage surges, loss of control, structural failure, or other action. Of these causes, arc-back is generally determining as it imposes a basic limitation on the rectifying action involving fundamental processes in the arc discharge. The limitation in rectifier capacity caused by arc-back is different in nature from the thermal load limitations of most electric machinery in that it usually results in an impairment in reliability, rather than any permanent electrical or structural damage. Furthermore, it is somewhat indefinite because of the random occurrence of arc-back and the complexity and lack of full understanding of the arc processes.

Temperature rise does not provide a measure of the loading on a mercury pool tube or indicate its probable life. However, the operating temperature of the tube does affect its capacity, as it controls the mercury vapor pressure and establishes conditions for the arc discharge.

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Thermal limitations of essential components, such as seals, studs, grids, also may affect rectifier capacity, although such limitations are usually readily diagnosed and easily surmounted.

As with most electric apparatus, time is an important factor in rectifier capacity, since the short time capabilities are different from those for continuous operation at a steady sustained load. The time required for a mercury pool tube to reach final temperatures is usually short because of the low thermal capacity of its parts. Furthermore, the time required to establish conditions favorable to the occurrence of arc-back involves all the factors affecting the arc action, of which temperature is but one. For these reasons the relation between magnitude and duration of the limiting load of rectifier tubes differs materially from that for other types of conversion equipment.

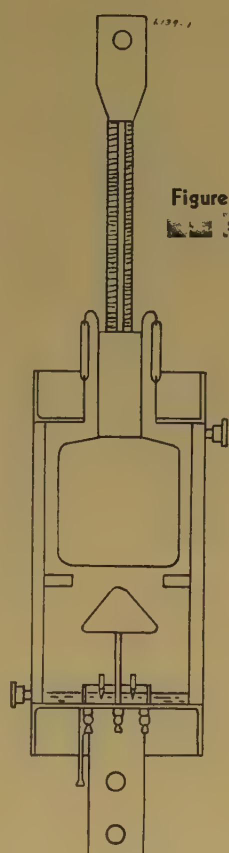

Figure 1. Cross section of [illegible] sealed ignitron

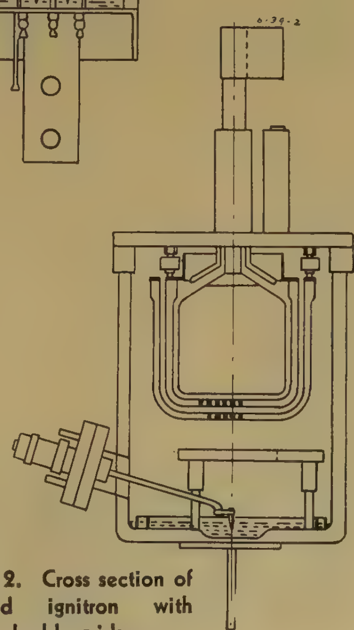

Figure 2. Cross section of pumped ignitron with double grids

Definition of Loading

The loading on a rectifying device may be described in terms of the magnitude and wave form of the current and voltage impressed upon it during a cycle. The loading is the sum of the effects of both separate and concurrent action of current and voltage. Inasmuch as the rectifying device performs only a switching action, the magnitude and wave form of the current and voltage depend almost entirely on the power circuit. The loading imposed upon the rectifying device by the circuit is termed the circuit duty.[2]

The capacity of a rectifying device may be expressed in terms of the circuit duty that it is capable of withstanding without failure. Some of the more important elements of the circuit duty are as follows:

1. Voltage—peak inverse and forward voltage; initial inverse voltage.
2. Current—average, peak, and rms current; final commutation rate.
3. Frequency (of a-c system).
4. Duration of loading.

Circuits differ in the amount of duty they impose upon the rectifying device because of differences in magnitude and form of the current and voltage waves. Also, the various elements of circuit duty have different weightings relative to the over-all loading on the rectifier. Since the effects of the various elements of circuit duty upon rectifier capacity are to a large degree mutually dependent, it is necessary to state the rectifier capacity for several different types of circuit duty in order to fully describe its capabilities.

It is often more convenient to state the rectifier capacity in terms of factors more directly related to the circuit constants and performance characteristics than in the elements of circuit duty. Some of these factors are:

1. Mode of operation (number of rectifier phases).
2. Direct voltage.
3. Direct current.
4. Amount of phase control.
5. Reactance.

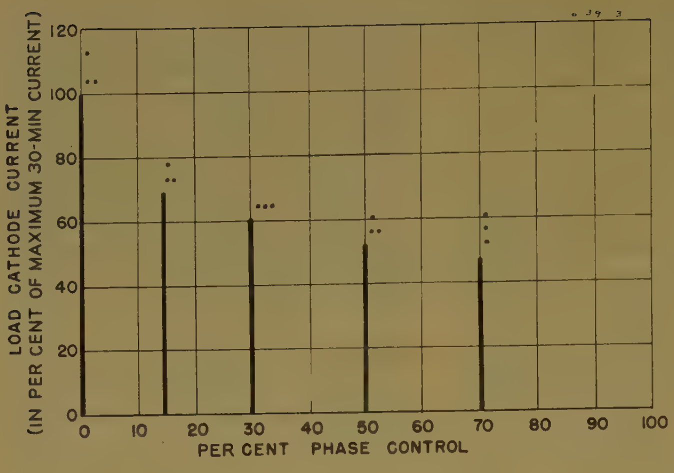

Figure 3. Results of 30-minute load limit tests on six sealed ignitrons

Dots indicate arc-back

A full statement of rectifier capacity also must include the capabilities and limitations under loadings of one cycle duration, that is, the fault current capacity. This capacity is defined as the crest value of current which the rectifying element can carry without incurring voltage surges during the conduction period or arc-back in the succeeding inverse period.

Determination of Capacity

The design of mercury arc rectifiers is based to a large degree on experimental data obtained from tests. As a consequence, each new design requires extensive tests to check its adequacy and determine its capabilities.

Rectifier capacity is determined from load test which consist in operating the rectifier under loads of increasing severity until failure occurs.

Load tests may be made on either a complete unit with a full complement of tubes, or on a section with only enough tubes to obtain the desired circuit action. Accelerated load tests have been termed load limit tests, and the procedure for making them has been described in a previous paper.[3]

Typical load limit test data are presented for the two ignitron tubes shown in Figures 1 and 2. Figure 1 shows a sectional view of a sealed ignitron and Figure 2 illustrates a sectional view of a pumped ignitron with double grids. Load limit data obtained with six sealed tubes connected to a double-Y transformer are shown on Figure 3. Similar data for three pumped ignitrons connected to a single Y are given in Figure 4.

The fault current capacity is determined from short-circuit tests which may be made either on a complete unit or a single tube. The testing procedure consists in increasing the current on successive trials until failure occurs.[3] Typical fault current test results are shown in Figure 5, which gives data for a sealed ignitron.

Characteristic Curves

Rectifier capacity usually is presented in the form of load–time curves, with the load given in amperes and with the other factors determining the load entered as parameters. These curves may be drawn directly from the data obtained on load limit tests, as the usual procedure is to make tests at a given voltage and vary loading by changing the load current.

Load–time curves for the sealed ignitron are shown in Figure 6 and for the pumped ignitron on Figure 7. The load–time curves are plotted most conveniently on a semilogarithmic scale because of the short thermal time constant of the tube and the short duration of the accelerated load limit tests. Values for the 30-minute and 2-hour points on the curves are obtained from the 30-minute load limit tests,* which are made by increasing the load by a fixed increment every half hour. The 30-minute point is then the highest load which was carried successfully without arc-back, and the 2-hour point is taken as the load three increments below. The load–time curves are usually almost straight lines, so values for times longer than two hours may be estimated by extrapolation. Final thermal conditions usually are attained in less than four hours, and experience has shown that the continuous capacity of rectifier tubes does not differ greatly from their 4-hour capabilities. The load capabilities indicated by these curves are limited by arc-back. Where structural features are limiting, different load time relations may be obtained.

A load–time curve may be drawn for each condition of rectifier operation, such as voltage, phase control, reactance, temperature, mode of operation. A complete statement of rectifier capacity therefore requires several families of load–time curves. However, it is frequently desirable to indicate the effect of the other load factors directly. The relationships between some of these factors and rectifier capacity follow.

Voltage

A given mercury pool tube may be operated over a wide voltage range. The load capacity in amperes decreases as the

* Previously designated continuous load limit tests.

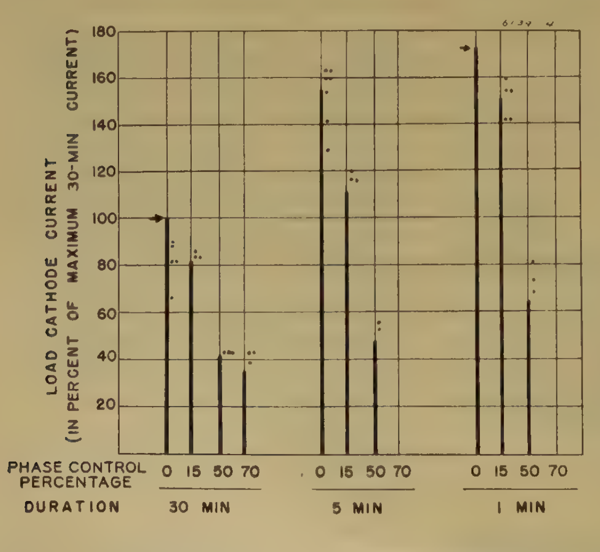

Figure 4. Results of load limit tests on three pumped ignitrons with double grids

Arrows indicate limit of testing facilities

Dots indicate arc-back

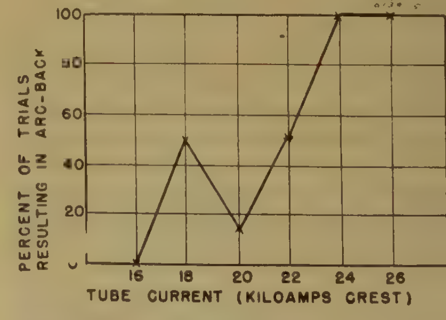

Figure 5. Fault current capacity of sealed ignitron

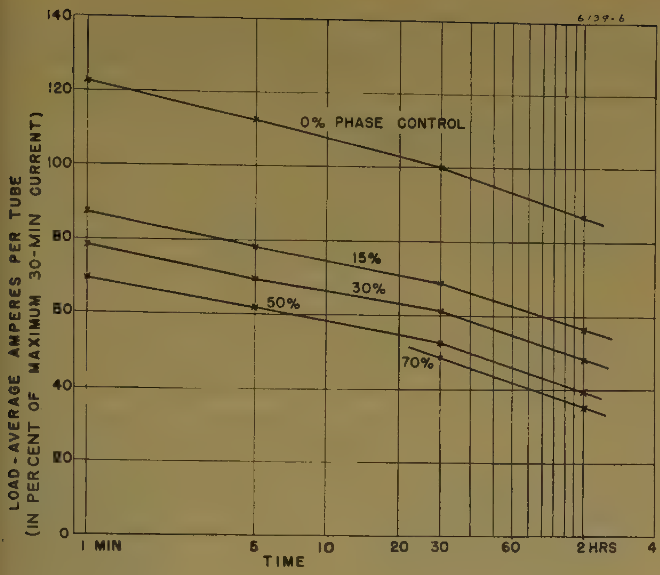

Figure 6. Load–time characteristics of a sealed ignitron

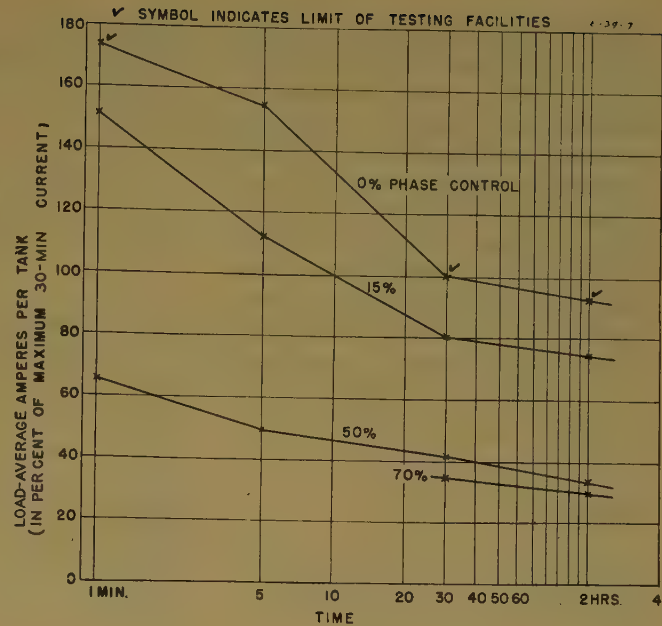

Figure 7. Load–time characteristics of a pumped ignitron with double grids

operating voltage is raised. However, the reduction in current capacity is usually less than the increase in voltage, in that the kilowatt capacity of a rectifier is generally increased as the operating voltage is raised.

These characteristics are shown best by plotting the rectifier capacity as a volt ampere relationship on a logarithmic scale, as in Figure 8. The volt ampere characteristics of both the sealed ignitron and the pumped ignitron are shown. Both have a greater slope than the constant kilowatt curve, indicating an increase in kilowatt capacity as the voltage is raised. Adequate data to establish the volt ampere characteristics as fully as might be desired are not available.

Phase Control

The effect of phase control upon the capacity of certain types of tubes is shown directly by the load limit data in Figures 3 and 4. These curves indicate a reduction in 30-minute current capacity when large amounts of phase control are applied. The percentage phase control is defined as the percentage reduction in output voltage.

Temperature

The rectifier usually is operated at its optimum temperature as determined by load limit tests. The operating temperature or temperature range, therefore, must be included with data on rectifier capacity.

Mode of Operation

The relative effects of the various elements of circuit duty upon rectifier capacity have not been established. Therefore, it is not possible to predict accurately the rectifier capacity for one mode of

Table I. List of Typical Tube Ratings

Tube Type and Description	Class of Service	Rated Output*: Kilowatts	Volts	Amperes
Type *A* ignitron	Industrial	150	250	600
		200	600	333
Type *B* ignitron	Industrial	300	250	1,200
		500	600	833
Type *C* ignitron	Industrial	8,100	9,000	900
Type *D* ignitron	Industrial	600	250	2,400
		1,000	600	1,666
Type *E* ignitron	Industrial	1,000	250	4,000
		2,000	600	3,330
Type *F* ignitron	Railway (Classes II and III)	4,000	3,000	1,333
Type *G* ignitron	Industrial	1,500	250	6,000
		3,000	600	5,000

* Rating of six tubes in double-Y circuit.

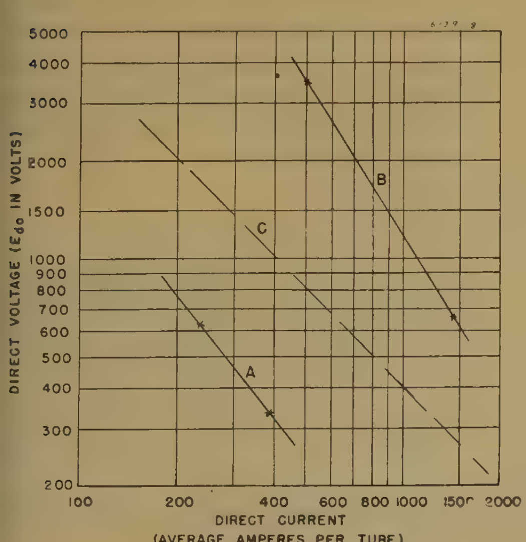

Figure 8. Volt ampere characteristics

A—Sealed ignitron
B—Pumped ignitron with double grids
C—Constant kilowatt

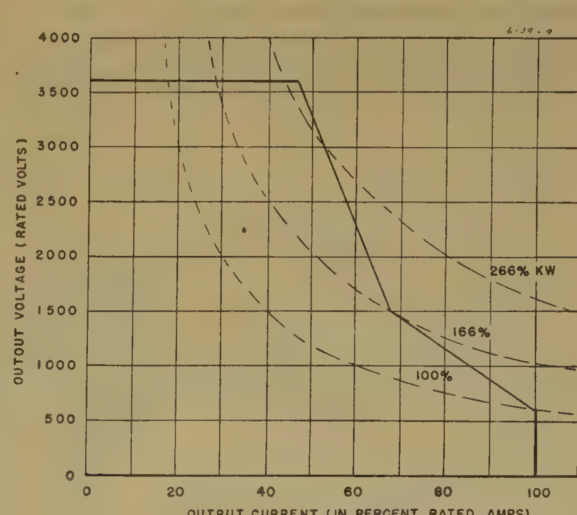

Figure 9. Typical rating curve for pumped ignitron—6-tube unit for railway service, classes II and III

operation from data obtained for another. Most rectifier applications employ circuits having the mode of operation of the double-Y circuit, and tests usually are taken using this circuit.

Research has indicated the importance of the arc phenomena occurring during the initial part of the inverse cycle.[4] The mode of operation, voltage, current, reactance, phase control, and frequency are all factors entering in this action. Our knowledge of this phase of rectifier behavior is not sufficiently complete to permit a complete treatment of these factors at this time.

Application

In the application of data on rectifier capacity as obtained from load limit tests to the design of a rectifier unit a margin must be allowed between the rated or anticipated load and the capacity indicated by tests. This tolerance must be provided to cover the following contingencies:

1. Differences between normal and accelerated loading. It must be recognized that load limit tests are short time or accelerated tests and do not duplicate accurately all the effects of sustained operation in service.

2. Differences in circuit duty. Where load limit tests are made on a set of three tubes, the circuit duty may be less than that incurred on a complete rectifier circuit.

3. Transformer unbalance. Unbalanced transformer reactances may cause unequal tube currents. Transformers usually are designed to permit an unbalance in neutral currents not exceeding five per cent. (The per cent unbalance is defined as the ratio of the difference between actual and rated neutral current to rated neutral current.)

4. Service requirements. The requirements of various applications differ with respect to reliability, depending upon the economic penalties or operational hazards incurred by the occurrence of arc-back.

The amount of the margin will depend upon judgment and experience involving the above factors.

Sizes and Ratings

Table I gives a list of typical tube ratings for some of the ignitron tubes now available.

As the same tube may be used over a wide range of voltages and may be applied to different kinds of service, such a listing does not completely describe its capabilities.

A fuller description of the tube capabilities is provided by rating curves of the type shown in Figure 9.

Conclusions

A truly comprehensive statement of tube capabilities directly in terms of the elements of circuit duty is recognized as an ultimate objective. This objective is not fully possible with our present knowledge.

However, a procedure for the description of the load capabilities and limitatations of mercury pool tubes has been outlined, and a convenient and useful form for the presentation of such information has been developed. The adoption of a standard form for the specification of tube capacity will help application and operating engineers prepare specifications for tubes which are to be used in rectifier equipment.

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Mica Capacitors for Carrier Telephone Systems

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Synopsis: Silvered mica capacitors, because of their inherently high capacitance stability with temperature changes and with age, now are used widely in oscillators, networks, and other frequency determining circuits in the Bell Telephone System. Their use in place of the previous dry stack type, consisting of alternate layers of mica and foil clamped under high pressures, has made possible considerable manufacturing economies in addition to improving the transmission performance of carrier telephone circuits. These economies are the result of their relatively simple unit construction and the ease of adjustment to the very close capacitance tolerance required.

PRESENT DAY carrier telephone systems are placing special demands on capacitors used in frequency determining and frequency sensitive circuits with particular emphasis on initial precision, low a-c losses, and high capacitance stability with temperature and time. In these systems mica capacitors are used in the oscillating circuits and also are used extensively in the electric filters employed to segregate the numerous and closely spaced carrier frequency channels. For these applications the transmission potentials applied to capacitors are quite low and in general do not exceed a few volts. These capacitors, however, must meet very stringent requirements for capacitance tolerance at the time of manufacture, and any subsequent changes in capacitance caused by temperature changes and age must be held to extremely low values to insure high quality transmission for the systems. As a result considerable effort has been expended by the Bell Telephone Laboratories and the Western Electric Company in the development of designs of mica capacitors which would lend themselves to large scale economical production, and at the same time provide high initial precision of capacitance and high capacitance stability over long periods of time under service operating conditions. To meet these exacting demands, mica capacitors having sprayed silver electrodes intimately bonded to the mica laminations were developed and introduced several years ago into the Bell System in quantities to replace the former less stable and more expensive designs.

As a matter of historical interest, silvered mica capacitors were manufactured by the Western Electric Company and used in the Bell System more than 40 years ago as capacitor standards in laboratory measurement work. These silvered mica capacitors were made by the relatively expensive method of chemical deposition of the silver on the mica.

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A. J. CHRISTOPHER, supervising engineer on capacitor design, and J. A. KATER, capacitor design engineer, are with Bell Telephone Laboratories, Inc., New York, N. Y.

Physical Construction

Figure 1 illustrates the former type of construction known as the dry stack mica capacitor. In this construction alternate layers of mica and tinned copper foil are held together under high pressure by a clamping arrangement. The unit is maintained in a sealed and nonimpregnated condition by potting in a metal container in a viscous asphaltic compound which does not penetrate the unit appreciably even in its fluid state.

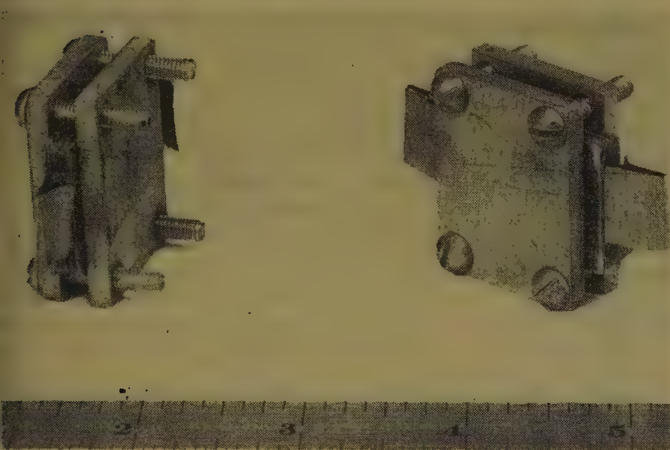


Figure 1. Dry stack mica capacitor units

Two distinct types of silvered mica capacitors were developed for general use in carrier telephone systems. One type known as the potted design was provided for applications where the highest initial precision and highest capacitance stability are required for services involving a limited range in operating temperature usually from about 60 to 100 degrees Fahrenheit. These capacitors are provided in several sizes of extruded nonmagnetic containers and provide for any desired capacitance value up to 0.4 microfarad. The other type is provided with mica filled molded bakelite casings and was developed for applications where small physical size is essential, but where the higher initial precision and higher capacitance stability of the potted types are not required. This type is provided in three physical sizes to cover a capacitance range from a few micromicrofarads up to approximately 20,000 micromicrofarads. A more detailed description of the construction of these two types of capacitors will be given.

Figure 2. Silvered mica lamination

Figure 2 shows the area of the mica lamination covered by the silver electrode. This area is coated with a suitable silver solution by spraying after masking the areas required for margins. The laminations then are fired at a high temperature for a short period to form a tight bond between the silver and mica. The laminations then are stacked, as shown in Figure 3, with a continuous strip of lead foil F interleaved between silvered surfaces so that one set of silver electrodes is joined electrically together and to the foil. Another strip of foil likewise is connected to the other set of electrodes on the opposite faces of the laminations. The purpose of looping-in the foil instead of using separately laid-in strips of foil is to permit a continuous indication of capacitance during stacking by means of a precise capacitance meter connected to the ends of the foil. Adjustment for capacitance first is made roughly by adding a sufficient number of laminations so that the measured value is somewhat larger than required. Terminals T then are crimped tightly over the ends of the foil to the mica stack to complete the unit assembly. The assembled unit of the potted capacitor after drying and coating with a thin layer of mineral wax, then is finally adjusted to close tolerance by removing the silver coating as required from the upper exposed surface shown in Figure 3. This has made possible the provision of capacitors meeting commercial tolerances of $\pm(0.2\%+2$ micromicrofarads) after being potted in wax in metallic containers. The molded case capacitor units are treated and adjusted in the same manner except that the wax dipping is omitted and the final adjustment prior to molding is accomplished by removing silver coating from the top lamination in a dry condition. Because of less predictable capacitance shift caused by molding the best commercial tolerance for molded capacitors is $\pm$ (0.5 per cent + 1 micromicrofarad). To insure high over-all quality

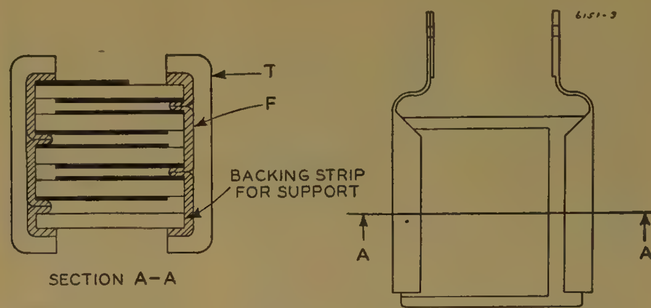


Figure 3. Silvered mica unit construction

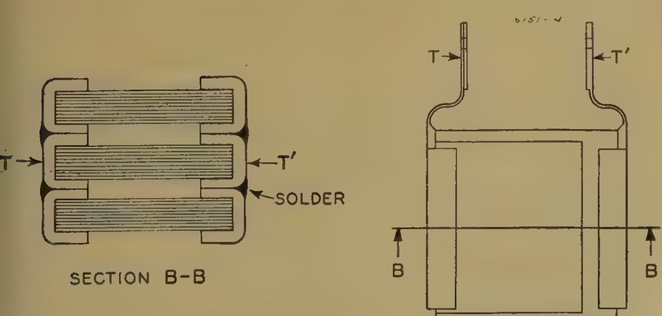


Figure 4. Multiple unit construction

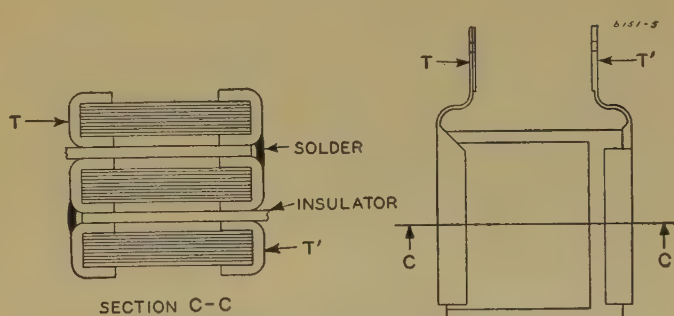


Figure 5. Series unit construction

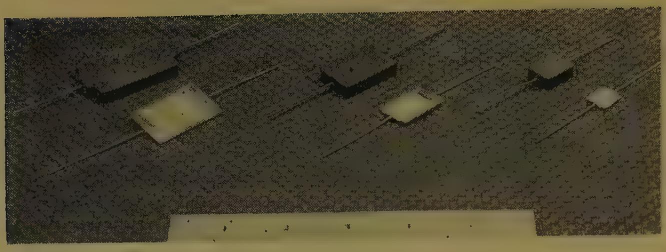
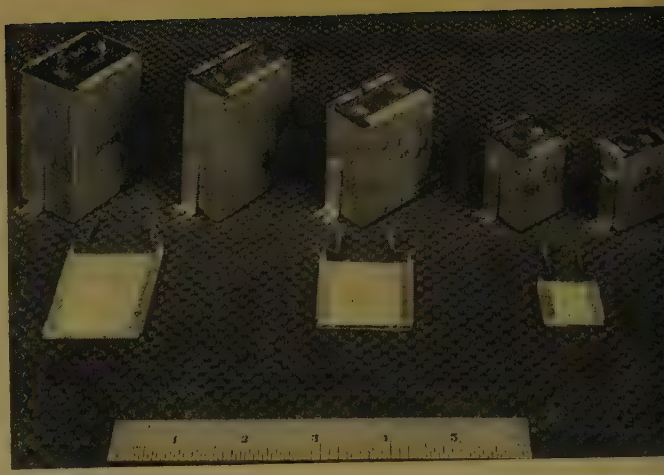


Figure 6. Molded silvered mica capacitors in current production

Figure 7 (right). Potted silvered mica capacitors in current production



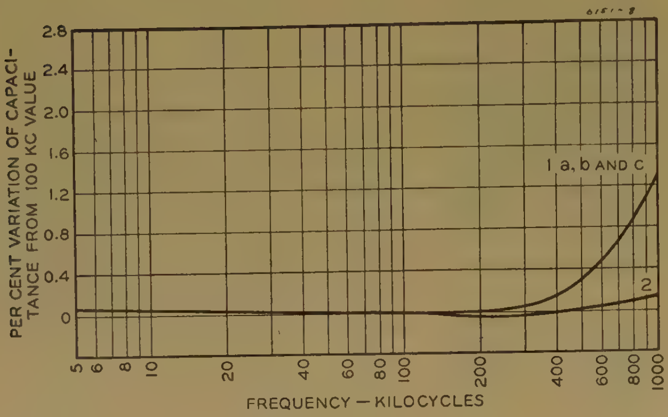


Figure 8. Variation of capacitance with frequency

1a = potted silvered mica, 0.01 μf
1b = molded silvered mica, 0.01 μf
1c = potted dry stack mica, 0.01 μf
2 = same as 1, except 0.001 μf

and performance all assembly operations are carried out in air-conditioned space where the relative humidity is maintained under 40 per cent at 75 degrees Fahrenheit.

Two or more of these unit assemblies are joined together mechanically and electrically to provide for the higher capacitances for the potted designs, as shown by Figure 4. For applications, such as line filters where high surge voltages may be encountered, several sections are connected in series, as shown in Figure 5.

In the case of the molded design, only single units are employed in the assembly because of the high molding pressures. The three sizes are shown in Figure 6. Figure 7 shows the five sizes of the wax potted designs in extruded containers which are provided for use in the Bell System plant.

Electrical and Performance Characteristics

The following discussion will deal with the more important electrical characteristics of dry stack and silvered mica types of capacitors and will cover data obtained on commercially made samples considered to be typical for these two types. Figure 8 shows the variation of capacitance with frequency over a frequency range of 5 to 1,000 kc for potted dry stack, potted silvered mica, and molded silvered mica capacitors of 0.01 and 0.001 microfarad capacitance. The capacitance change shown up to about 200 kc for the large capacitance value and up to about 400 k for the smaller value is dependent on th dielectric coefficient of the mica itself. I will be noted that this change is within 0. per cent. At higher frequencies th capacitance increases because of the effec of series inductance mainly contribute by the terminals. In the case of capaci tors intended for operation at high fre quencies, the effect of this inductanc must be taken into account and prope allowance made for it where high capaci tance precision is required.

Figure 9 shows the change in Q wit frequency of 0.026 microfarad dry stac and silvered mica potted capacitors fro 1 kc to 100 kc, and Figure 10 shows th change in Q of 0.001 microfarad potte and molded silvered mica capacitors fro 5 to 1,000 kc. The average Q of the si vered mica potted design is higher, and th variation of Q with frequency is somewha less than that of the dry stack design. I is believed that the inherently higher Q o of the silvered mica design results fro reduced losses caused by the eliminatio of asphaltic potting compound and elim nation of absorbed hydrocarbon materia as well as small traces of water vapor fro the surface of the mica by the very hig

Figure 9. Variation of Q with frequency for potted silvered and dry stack mica capacitors

1 = silvered mica, 0.026 μf
2 = dry stack mica, 0.026 μf

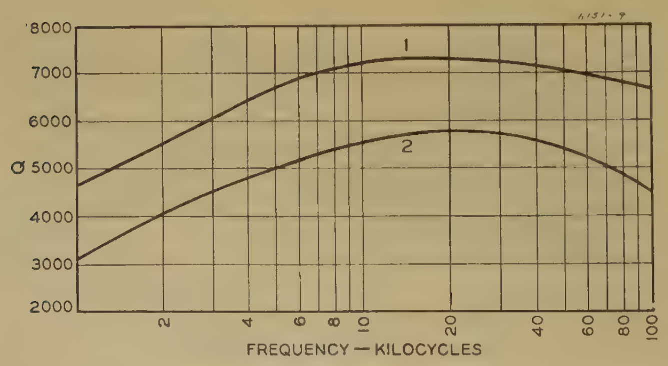
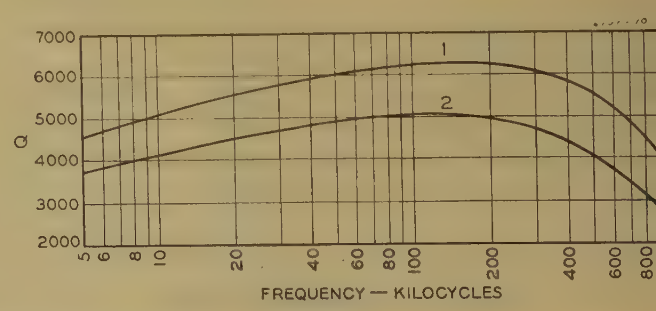


Figure 10. Variation of Q with frequen for potted and molded silvered mica capacit

1 = potted silvered mica, 0.001 μf
2 = molded silvered mica, 0.001 μf



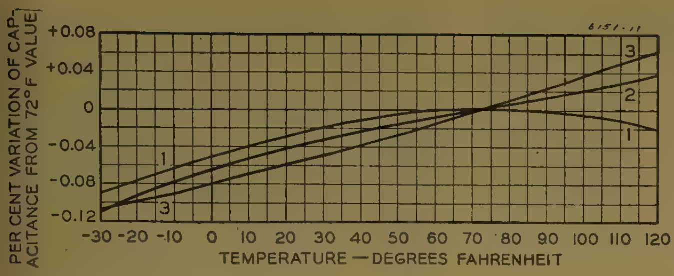


Figure 11. Variation of capacitance with temperature

1 = potted silvered mica, 0.01 μf, 10 kc
2 = molded silvered mica, 0.01 μf, 10 kc
3 = potted dry stack mica, 0.01 μf, 10 kc

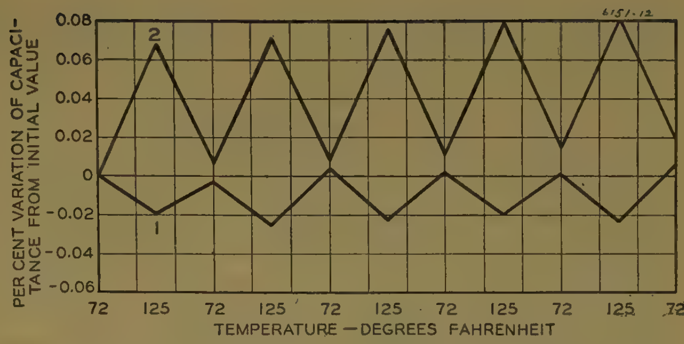


Figure 12. Variation of capacitance with temperature cycling

1 = potted silvered mica 2 = potted dry stack mica

firing temperature. The Q of the potted silvered mica also is higher than that for the molded capacitor of the same capacitance value. The dielectric losses in the phenolic casing and some contamination of the unit from the liberation of vapors during molding probably account for the lower Q of the latter.

Typical changes of capacitance with temperature for the molded and potted types are shown in Figure 11. It illustrates a pronounced difference between the molded and dry stack potted types as compared to the potted silvered mica construction. In the case of the molded and dry stack potted types, the change of capacitance with temperature primarily is caused by changes in the physical dimensions of the unit. These changes are such as to produce a positive temperature coefficient throughout the entire temperature range. In the case of the wax treated and wax potted silvered mica capacitors, the capacitance changes with temperature are caused by changes in the physical dimensions of the unit and also by the effect of the wax, having a high negative temperature coefficient, on the fringing capacitance of the unit. As a result the temperature coefficent is positive up to about 60 degrees and is negative above approximately 80 degrees. Between 60 and 100 degrees, the usual operating temperature range for central office installations, the temperature coefficient is extremely low.

Figure 12 shows the variations in capacitance of a typical potted silvered mica capacitor and also, for comparison, the same characteristic for a typical dry stack potted capacitor with repeated changes in temperature from room temperature to 125 degrees. The drift in capacitance resulting from repeated temperature cycles is caused by such conditions as small permanent displacement of the laminations in the assembly, small changes in the effective separation of the electrodes, and possibly small changes in the effective dielectric constant. Figure 12 illustrates how the drift in capacitance caused by the foregoing conditions, is minimized in the silvered mica capacitors. Data indicate that the potted silvered mica capacitors, also are more uniform from sample to sample in over-all capacitance stability, the extreme variations being one-fourth to one-fifth the magnitude of the variations observed with the dry stack design over the usual central office temperature range. In the case of the molded capacitors, while the average capacitance changes caused by drift as well as temperature coefficient are less than for the dry stack type, their capacitance changes have been found to be approximately double those of the potted silvered mica designs under the same conditons.

Figures 13 and 14 show the third order modulation characteristics for the potted and molded types. This characteristic is a measure of the capacitor's creation of new and unwanted frequencies while transmitting frequencies impressed upon it. In the case of capacitors, the creation of unwanted frequencies is the result of nonlinear effects caused by mechanical vibration and dielectric hysteresis. Un-

Figure 13. Variation of modulation with frequency

1 = potted silvered mica
2 = molded silvered mica
3 = potted dry stack mica

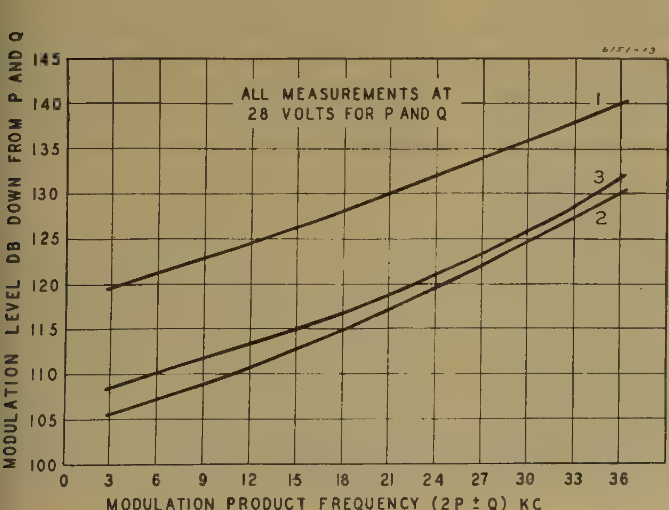
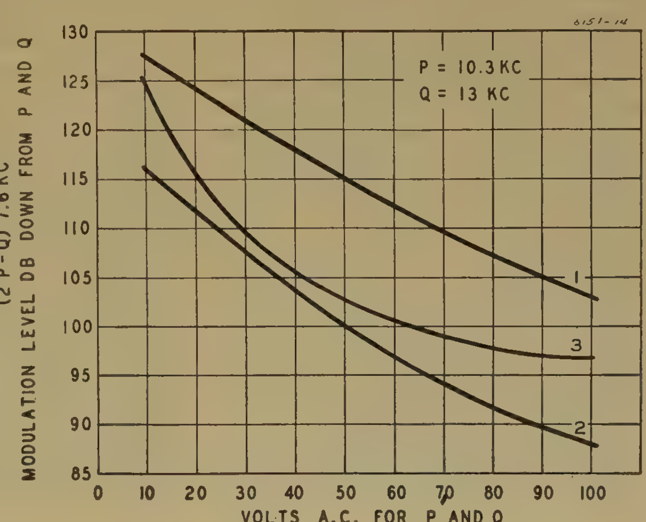


Figure 14. Variation of modulation with applied voltage

1 = potted silvered mica
2 = molded silvered mica
3 = potted dry stack mica



less these unwanted frequencies are held to very low energy levels as compared to the energy levels of the desired frequencies, in some cases less than one part in ten billion (100 decibels), crosstalk between carrier channels or noise or both seriously may impair high fidelity transmission of carrier telephone systems. It will be observed that over the frequency and voltage range shown, the silvered mica potted capacitors are approximately 15 decibels better than the molded designs. It is believed that this is because of the damping of mechanical vibration by the wax in the potted designs.

Conclusions

Current practice in the design of carrier telephone and other communication systems is imposing special requirements on capacitors used in frequency sensitive circuits. The exacting requirements of wire communication usually are not encountered to the same extent in other fields of communication engineering. The multiplicity of channels on a single pair of conductors, the large number of elements involved in a long circuit, and the constant use of these channels for years without adjustment or replacement of circuit elements impose very stringent requirements on capacitance tolerance at the time of manufacture, and on permissible change in capacitance subsequent to manufacture, to insure high fidelity transmission of voice and similar signals.

As has been described, two types of silvered mica capacitors were developed to meet these requirements. One type is wax sealed in a seamless metal container and has a practically flat capacitance temperature characteristic over the operating temperature range most commonly encountered, low modulation, and greater uniformity in the commercially manufactured product. The other type is provided in a low loss molded casing and satisfies a demand for small size and low loss capacitors, primarily for application where the higher precision and stability of the silvered mica potted types are not required.

A New Design for the A-C Network Analyzer

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IN THE PAST FEW YEARS, with the heavy demands made on power systems, an increased interest has developed in network analyzers for public utility use. This has resulted in installation of several new analyzers and design or planning of others. In line with this trend, the Iowa State College has been active in developing and constructing an analyzer for use by utilities of the midwest area. Following general educational policy, the work was undertaken as a research project of the Iowa Engineering Experiment Station to determine the feasibility of applying new equipment and new techniques in improving the design of the a-c network analyzer.

In beginning the work, investigations were made which established a major premise, namely, that higher frequencies than those employed in the past could be used profitably. However, if frequencies are carried higher than previously, conventional instruments become inapplicable, and it is necessary to adopt as a corollary to the major premise, that complete electronic instrumentation will be desirable. As an additional advantage for this system, the instrument burden on the network can be reduced to a desirable minimum, and no longer serves as a major design limit on currents, voltages, and impedances employed.

One reason for raising frequency was a desire to draw on the techniques and apparatus common to electronics and radio, to determine if an amalgamation of the power and radio fields might lead to an improved design. Power levels used in network analyzers are approximately those encountered in electronic equipment. Therefore, much of the standardized high production equipment for electronic service might be incorporated profitably in a network analyzer, instead of using scaled down power machinery and techniques. An example of the advantage of electronic equipment was found in the design of electronic generators for an analyzer. The electronic generator can be built with performance improved over that of the conventional induction motor phase shifter type, and at a much reduced cost.

An important economic factor influencing the decision to operate at a higher than normal frequency was the possibility of lowering cost through use of smaller values of inductance and capacity for given values of reactance. A valuable technical advantage which could be realized at a higher frequency was the elimination of iron cores in the inductive reactors, with consequent improvement in constancy of inductance at all current values and freedom from wave form distortion.

It should be understood that while the network analyzer to be described has proved its ability and the correctness of the fundamental design assumptions, it will continue in an experimental state for a considerable period of time. Built as a small pilot model, to prove the design, and to handle only simple problems, the analyzer may be increased in size by the addition of units in the future. Some modifications based on experience may be necessary, and compromises in equipment choice, forced by wartime conditions, may be eliminated in the larger model.

Fundamental Design Features

In the design of an analyzer in a new frequency range, extrapolated far beyond previous experience, many design decisions had to be made with and without adequate analytical background. Most of the decisions were interdependent and not simply related, but an attempt has been made to give the basis of many of them in the material below.

Choice of Frequency

Since all units are calibrated in terms of resistance, reactance, or susceptance, the actual frequency chosen for operation is independent of the frequency of the system being studied. Previous designs of analyzers have been operated in the 400 to 500-cycle range. In view of certain obvious advantages from the cost standpoint, a higher frequency seemed a desirable choice. These economic advantage

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are twofold. Raising the frequency reduces the size of inductors and capacitors required for given reactance values, thus reducing the cost. Also, if the frequency is raised sufficiently, the ratings of many parts become those common to the radio industry, and the high production in this industry further reduces the price. Radio industry components are made in a great variety of sizes and are easily obtainable, which are both desirable advantages. While of somewhat variable quality, it is possible to obtain radio parts which are entirely dependable. This can be accomplished by careful selection of the manufacturer and avoidance of parts obviously built down to a price.

The curve shown in Figure 1 illustrates the variation in price of 100 ohms of capacitive reactance at various frequencies. They are list prices for good quality capacitors, and it is assumed that units of the exact required capacitance could be purchased with no paralleling of small units being necessary. It is probable that inductor costs also would show a considerable drop with frequency, although, as inductors are not common items of commerce, comparable prices cannot be obtained. The curve shown in Figure 1 is not meant to apply to capacitors alone, but is intended to give an indication of the trend in both inductor and capacitor costs with frequency.

Based on this reasoning, a tentative selection of the frequency base for the analyzer was made. This selection was subject, of course, to confirmation or change as other design features were checked. The frequency chosen for the Iowa State College a-c network analyzer was 10,000 cycles per second. This represents a major extrapolation of previous experience. Choice of a frequency substantially below 10,000 cycles per second would have made inclusion of iron cores in the inductors a necessity, and choice of a frequency much higher would have led to difficulties caused by reactance and susceptance of interconnecting leads, as will be discussed. An additional factor in the choice of a frequency of exactly 10,000 cycles per second was the commercial availability of a satisfactory source of low cost.

Frequency Source

As the analyzer was to be operated from a small power system of rather unstable frequency and because a rotary machine of small power rating at 10,000 cycles with sine wave form might be difficult to obtain, all-electronic operation of the analyzer was decided upon. This decision meant the use of a temperature-controlled quartz-crystal standard-frequency oscillator, operating at 100,000 cycles per second as the primary frequency source. By occasional checking against standard frequency signals from radio station WWV of the Bureau of Standards in Washington, this oscillator can be maintained within one cycle at 100,000 cycles, an accuracy in frequency of 0.001 per cent. In actual service, without adjustment, the error has never been found greater than five cycles at 100,000 cycles.

With frequency dividing circuits, the 100,000-cycle signal is subdivided by a factor of ten, yielding a 10,000-cycle output of a few volts, with reasonably good wave form, as the primary frequency source for the analyzer. This voltage then is filtered for wave form improvement and applied to amplifier circuits which are used to develop the power needed in the outputs of the equivalent generating stations of the analyzer.

Choice of *E*, *I*, and *Z* Base Values

With metering limitations somewhat removed in setting base values of voltage, current, and impedance, various other factors to be considered are physical size of units, and electric and magnetic couplings. Also needing consideration are the resistance and reactance drops and susceptance currents in leads connecting the units to busses and transfer jacks.

A low impedance base decreases inductor reactance and reduces the effects of stray shunt capacities. Shunt capacitor ratings are increased, but, in the radio capacitor field, physical sizes are so standardized that an increase in rated capacity over a ten to one range may mean no appreciable change in physical size or cost. Capacitor sizes, and expected effects of shunt lead capacities, indicated that a 100-ohm impedance base would be desirable and it was adopted. To keep electric coupling between circuits within units to a minimum, a voltage of ten volts was selected, resulting in a base current of 0.1 ampere. The power dissipated in the network is so low that small resistors may be used safely without fear of burning out on accidental overloads.

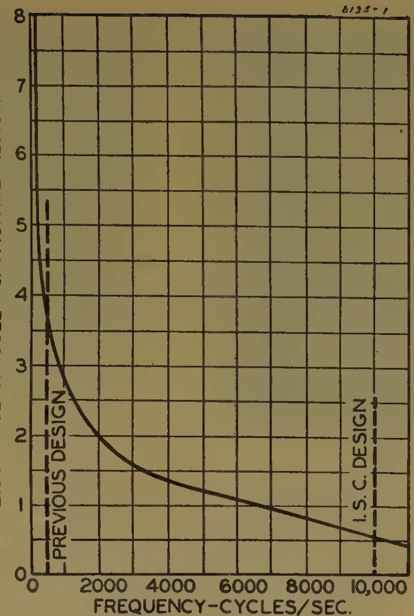


Figure 1. Estimated cost of 100 ohms of capacitive reactance at various frequencies

Interconnection of Units

For accurate operation of the analyzer, all leads from units to busses, transfer jacks, and so forth, should introduce negligible resistance, reactance, and susceptance into the circuit. At 10,000 cycles the problem of inductive and capacitive coupling between circuits can be severe unless transposition or shielding is used. Various types and sizes of wire for these lines were investigated with respect to resistance, series reactance, and current taken by the susceptance. Standard tolerances were set up as 0.1 per cent of base values.

Calculations were made on standard types of twisted pair lines used for radio frequency transmission, which indicated that certain types would meet the specifications on resistance and reactance, but might cause larger than desired shunt capacity currents. It was hoped that the twisted pair feature would introduce sufficient transpositions to reduce coupling. A better solution for elimination of intercircuit coupling was the coaxial cable, and computations indicated that some types would have satisfactorily low circuit values. The complete absence of external field eliminates all circuit coupling, and a $^3/_8$-inch diameter double conductor (Twinax) coaxial cable was adopted for interconnections. The double conductor and shield feature allows input, output, and common ground leads of a line, load, or other unit to be carried in a single cable. The capacity between the two wires then appears shunted across the line or load circuit elements. The error introduced by this capacity across the circuit elements is 0.25 per cent of base value, for the longest lead encountered. The capacity of the two wires to the shield and ground is negligible, causing a current flow of only 0.075 per cent of base current on the longest lead. The resistance of the chosen cable for the longest length is approximately 0.1 per cent of base value.

The reactance of the chosen Twinax line was another limiting factor and for

Figure 2. General view of the Iowa State College a-c network analyzer

the longest lead is 0.1 per cent of base value. This fact influenced the choice of 10,000 cycles as base frequency. The magnitude of these limiting factors indicates that a cable design with slightly larger conductors would be desirable. Such a cable was not a commercial product and under wartime conditions it was necessary to design around available equipment.

Inductive Reactances

One of the reasons for the choice of 10,000 cycles as base frequency was the possibility of eliminating all iron in the line and load inductors by the use of air core coils. Use of air core coils removes all anxiety as to wave form distortion by the iron, and constancy of inductance at all current values. Lower losses or higher ratios of reactance to resistance also have been achieved than those mentioned for iron core coils in lower frequency network analyzers.

Air core coil designs have been developed for all sizes of reactors used, which give all inductors phase angles of above 88 degrees, and certain much used load inductor values above 89 degrees. All coils are shielded in mutually perpendicular pairs, and the above phase angle measurements are with shielding in place. The resistance component introduced by any coil of a load unit may be considered as two per cent of the rated reactance, and of any coil of a line unit as four per cent of rated value. Each decade of a reactor group is obtained by series switching of 1-, 2-, 3-, and 4-ohm reactors or multiples of these values.

Accuracy

A definite statement of the accuracy to be expected from such a complex mechanism means very little because of additive and subtractive errors introduced by so many circuit elements in cascade, and to usage of meter readings in various types of calculations. It can be stated that all resistor elements are accurate to better than one per cent of the setting. All inductive reactors have been adjusted to within 0.2 per cent and, because of the air core construction, will maintain that accuracy at all current values. Capacitive susceptors have been assembled to match the 0.2 per cent maximum deviation requirement. The metering system has been calibrated from d-c standards by thermocouple transfer units to one-half per cent or better of full scale values for normal loading conditions.

Use of Standard Radio Equipment

It already has been emphasized that the selections for frequency and impedance bases made possible the use of standard high production radio components and coaxial cable. In addition, the radio and telephone industries have developed a line of standardized instrument racks and panels which have advantages in flexibility and availability. These panels are 19 inches wide with heights in multiples of $1^3/_4$ inches. The racks are 20 inches wide, six feet high, with mounting holes so arranged that panels can be mounted at any point desired. Such an arrangement will permit easy changes in the future if needed. A view of the complete analyzer is shown in Figure 2.

Control Desk

The main control and metering desk, illustrated in Figure 3, was designed in modern style and finished to harmonize with the panel equipment. Its design permits the operator an easy view of the complete analyzer and houses all instrument amplifiers and certain relays.

The desk combines analyzer controls and master instruments on a sloping panel. The angle of the panel was chosen to be comfortable for the operator in either controlling the analyzer or taking data. On the panel, and accessible to the left hand, are the main 60-cycle power controls, lever keys for connecting the instruments to either input or output circuits of any analyzer unit, a watts-vars switch, and a switch to place the desk instruments in operation.

Also operated by the left hand are the push buttons for preselection of relays by which the desk master instruments may be connected in any circuit. By simply pushing buttons corresponding to the number of the unit to be metered, as on an adding machine, the relay circuits are selected. The capacity of the push button system is 999 circuits which, by reason of the input and output relays on each unit, will permit metering at 1,998 distinct points should that number ever be required. All push buttons are mechanically interlocked to prevent metering two circuits at once. By assigning these controls to the left hand, a pencil held in the right hand for taking data or other work need not be dropped during operation.

Figure 3. The main metering and control desk

Figure 4. Front panel of a generator unit

The dials are for adjusting voltage and phase angle

The desk instruments are subsurface mounted and internally illuminated with fluorescent lamps, giving a very uniform glareless shadow-free light. The upper and lower rows of instruments are placed at different vertical angles so that an observer of average height will have a line of sight perpendicular to the meter scale for both instrument rows. This materially reduces the fatigue of the operator. Mounted below the instruments are range switches for the ammeter and voltmeter and a reversing switch for the wattmeter. Working in conjunction with the range switches are colored signal lights above the voltmeter, ammeter, and wattmeter, indicating the range multiplier in service at any time.

The instruments used all are operated electronically and comprise a voltmeter, an ammeter, a watt-varmeter, and two phase angle meters. The phase angle meters are at each side in the lower row, the other instruments being placed in the upper row with the wattmeter in the center. Amplifiers for operating these instruments are installed inside the control desk. The voltmeter and ammeter are fairly conventional electronic amplifier types, stabilized by large values of degenerative feedback. The wattmeter is an electrodynamometer type with both current and voltage coils supplied from stabilized amplifiers. A phase shifting circuit converts this to a varmeter. The phase meters are an unconventional feature. One reads angle of the voltage at the metered point and the other reads angle of the current, both with respect to any desired common reference voltage. A switch on the desk allows selection of any generator electromotive force as a reference, or any desired point in the network may be connected as a reference through the transfer jacks. The phase meters employ an electronic circuit[6] which clips and squares the waves before phase comparison, eliminating error caused by amplitude variation.

Insertion of the instruments into the desired circuit is handled by relays, selected by the push buttons. The relays connect the circuits of the desired unit to a Twinax metering bus, which terminates in current shunts and voltage dividers, across which are connected the master instruments of the desk. The shunt introduces a resistance of 0.1 per cent of base value, while the voltage divider inserts a loading of 0.03 per cent on the circuit. The metering bus, connected to all units, inserts some distributed capacity, and the contacts of the switching relays in each unit

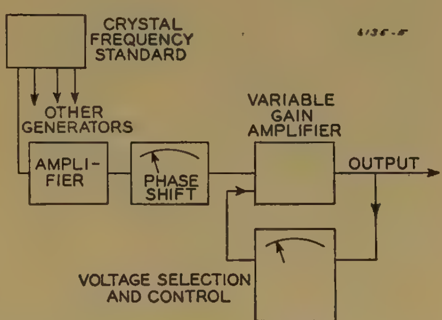


Figure 5. Outline diagram of the generator circuit and operation

insert large lumped capacities in the metering system. These all have been compensated for effectively and the resultant error is negligible.

Design of Units of the Analyzer

The present Iowa State College a-c network analyzer, although a small pilot model, was designed to handle most advantageously the simpler problems for the utilities of the midwest area. In consultation with these utilities, the following choice of numbers of units of the analyzer was made: 4 generator units, 24 line units, 12 load units, 12 capacitor units, 4 tap changer units, and 4 mutual transformer units. Features of the design of each of these units will be given.

Generator Unit

The standard frequency source has an output of a few volts and negligible power. Each unit simulating a generating station of the analyzer must amplify the voltage to the desired base level or higher and provide sufficient power for input to the loads and lines of a given problem. The generator also must have independent means of adjustment for phase angle and magnitude of voltage output.

The generators for this analyzer consist of vacuum tube amplifiers of power output considerably greater than required for rated base output. A generator is shown in Figure 4. Phase angle control is provided in two 180-degree steps through the left hand dial and the "plus-minus" switch on each generator panel, the control operating in the low level amplifier stages. Voltage selection is possible by the right hand dial on the generator panel. With selection of output voltage, an automatic voltage control circuit is set to adjust the amplifier gain and maintain the desired voltage. A block diagram of a generator is shown in Figure 5.

The performance of the automatic voltage control circuit is shown in Figure 6. The control operates to maintain constant voltage to 250 per cent of base or rated output current into a resistance load. This represents zero internal impedance and zero per cent regulation. Above 250 per cent load current, the generator voltage drops because of an internal impedance of about 4.0 ohms. On either leading or lagging load, the voltage regulation is improved over that for resistive load. Contrasting with the generator performance under control is the performance with no voltage control, as shown by the dashed line in Figure 6.

Connected to each generator is a set of instruments to read generated electromotive force, current, and output power, as shown in Figure 7. The instruments are electronically operated and stabilized by degenerative feedback in order to be entirely independent of changes in tube characteristics and line voltage. All in-

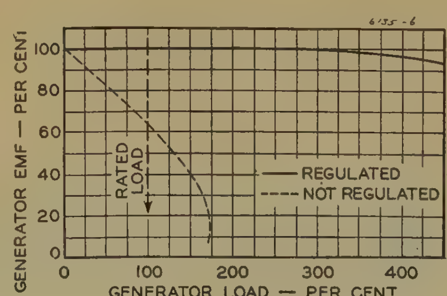


Figure 6. Voltage regulation of the generators for varying output

Figure 7. Individual generator metering panel

Figure 9 (right). Assembly of a line unit

struments are calibrated in percentage of base values. The wattmeter is arranged with zero center scale so that indications of reverse power flow can be read easily. With the meters permanently connected to each generator output, the loading can be checked continuously, and the effect of adjustments of one generator on the output of other generators in the system can be observed readily.

Mounted below each generator is an inductive reactance panel by which generator reactance up to 99-per-cent base value by 1-per-cent steps may be inserted. Both generator electromotive force and terminal connections are available at the transfer jacks for ready circuit connection.

Line Unit

In reducing an actual system to a model representation on a network analyzer, retention of each line of the system as an entity is highly desirable. In the majority of previous analyzer designs, the series resistance and reactance equivalents have been retained in the analyzer line unit, but shunt susceptance equivalents of all lines emanating from a common bus usually have been added and set up on these analyzers by a single separate capacitive susceptance unit. The direct metering of a line for terminal current conditions, therefore, was impossible. The line unit of the Iowa State College analyzer was designed as an entity containing the shunt capacitive susceptance as well as the series resistance and series inductive reactance. This detail was possible because of the favorable space and cost conditions inherent in the higher frequency design. A schematic wiring diagram of a line unit is shown in Figure 8, and a photograph of a line unit panel is shown in Figure 9.

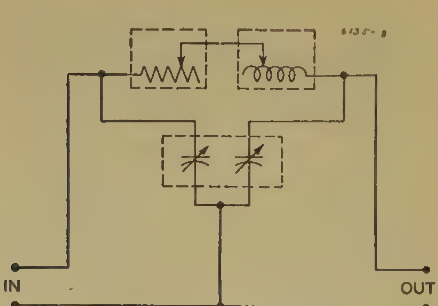


Figure 8. Simplified diagram of the wiring of a line unit

Either a T- or a π-section design would be electrically satisfactory for the line unit. However, the cost and space requirement of an average capacitive susceptance element is considerably less than that of an average inductive reactance and resistance element, and hence the π-section, requiring one set of reactance and resistance elements and two sets of susceptance elements, was chosen. In each line unit, the two sets of susceptance elements in the π-section are adjusted simultaneously by a 2-gang switch.

The series resistance and reactance are each adjustable from zero to 99 per cent in 1-per-cent steps. The total shunt susceptance is adjustable from zero to 9.9 per cent in one tenth of 1-per-cent steps. Each of these three elements thus was designed in two decades.

Load Unit

The basic elements in the load unit design are resistance and inductive reactance inasmuch as the majority of load conditions in an actual system may be represented by these parameters. In representing load conditions on the analyzer, often the voltage at the load point is an unknown quantity at the time of establishing equivalent resistance and reactance values. If a particular power and power factor are known conditions, an estimate of the voltage usually is made and the corresponding circuit parameters are calculated and set on the analyzer. Measurements then are made of the first approximation of the desired power. If the power is not that desired, a resetting of parameters is necessary. In the Iowa State College analyzer, the resistance and reactance values are viewed electrically through a variable ratio autotransformer, adjustable by 1-per-cent values from 1/1.10 to 1.10. If the first estimate of the resistance and reactance are within approximately plus or minus 20 per cent of the actual values required, the adjustment of the load may be accomplished quite simply by adjusting the turns ratio of the transformer, thus avoiding a recalculation of the individual resistance and reactance parameters. A schematic wiring diagram of a load unit is shown in Figure 10, and photographs of a load unit panel and rear assembly are shown in Figures 11 and 12.

The resistance and reactance elements may be placed either in series or in parallel with one another by the switching plan illustrated in the schematic wiring diagram. This permits more versatility in the application of this unit to a variety of problems.

The resistance and reactance are each adjustable from zero to 9,990 per cent in 10-per-cent steps. Each of these two elements was thus designed in three decades.

The output terminals of the load unit are retained even though for normal load usage this output is closed by a short-circuiting plug. By thus providing the output terminals, the load unit may be used also as a circuit of higher series im-

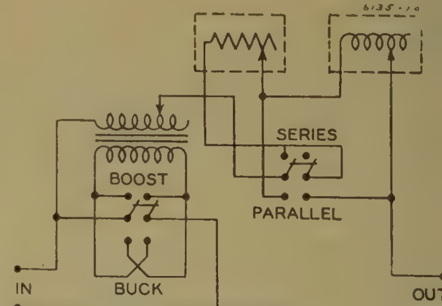
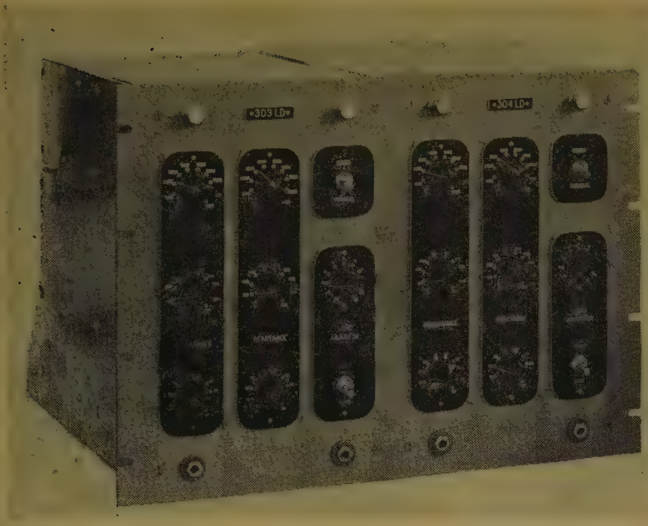


Figure 10. Simplified diagram of the wiring of a load unit



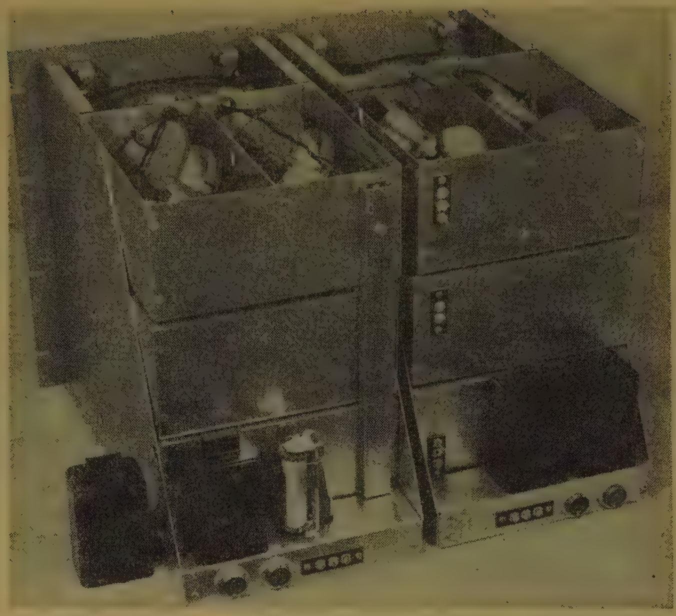


Figure 11 (above). Front panel view of a load unit

Figure 12 (right). Rear assembly of a load unit

pedance rather than a line unit for those applications where such a circuit may be desirable.

The autotransformer utilizing a high frequency low-loss steel core has an equivalent shunt impedance such that the error introduced by the adjusting transformer when used with the base impedance is approximately 0.6 per cent. The equivalent series impedance of the autotransformer when set at full ten per cent is 0.7 per cent. Each of these errors is comparable to the one-half-per-cent accuracy which was the design goal of this analyzer.

Capacitor Unit

Separate capacitor units are provided on the analyzer for increasing the range of the susceptance built into the line units, adding capacitive loading to the load units, or for representing series or shunt capacitors at any point in a system. A photograph of such a panel is shown in Figure 13.

The capacitive susceptance may be used either as a series or shunt element by a switching plan similar to that used in the load unit. The susceptance is adjustable from zero to 99 per cent in 1-per-cent steps. The element thus is designed in two decades.

Tap Changer Unit

Tap changer units are provided for representing networks which do not possess net unity turns ratios around closed loops. The unit is a variable ratio autotransformer adjustable from 1.000 to 1.295 by 0.005 steps. The transformer has a low-loss steel core and in design is similar to the load adjustor transformer of the load unit. The equivalent shunt impedance is practically identical with that of the other unit. Incorporated also is a reversing switch which changes the turn ratio from a step-up to a step-down basis.

Mutual Transformer Unit

A unity ratio 2-winding transformer with an inductive reactance placed across either side is used for representing mutual coupling between circuits of a power system. The mutual transformer, utilizing a low-loss steel core, introduces losses of approximately 0.4 per cent of base power if used across a 10-ohm reactor at full base current. The equivalent series impedance of the transformer is also 0.4 per cent and introduces negligible error of insertion.

Miscellaneous

The Twinax cable from each unit terminates at the transfer panels in two sets of 2-conductor jacks. One pair of duplicate jacks connect to the common and input conductor of the unit and the other pair to the common and output conductor. By the process of extensions from duplicate pairs, a large bus system may be established.

Associated with the input or output terminals on each unit are separate jacks which permit the interconnection of other units inside the point at which the meters are inserted by the relay system. Thus a single instrumentation of a composite group of units is possible when this is desirable, as for example if capacitive loading were added to a load unit by inserting a capacitor unit inside the metering point of a load unit.

Operating Experience

Although commercial operating experience of the analyzer as a whole is rather limited at the present date, such system problems as have been set up on the analyzer have yielded results which check at nodes and loops quite consistently within the design tolerances previously mentioned.

Several 2- and 3-generator problems which have been calculated have been set up on the analyzer, and measured values on these problems likewise have been within the design tolerances of the units and the instruments.

As elements of the analyzer have been completed over a period of the past 15 months, each has undergone extensive testing in the component form; and, where the experimental measurements did not conform to the design tolerance of one-half-per-cent accuracy, further modi-

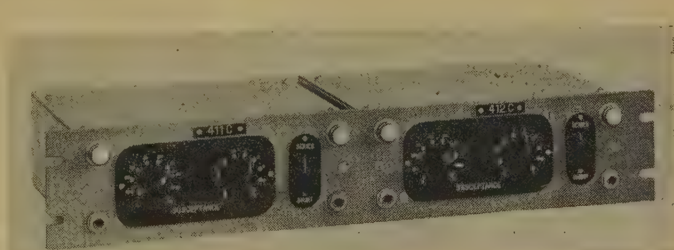


Figure 13. Panel view of a capacitor panel

A 400-Ampere Sealed Ignitron

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Synopsis: As rectifier capacity is increased, it becomes desirable to use larger capacity rectifying elements. The choice of rectifying element size requires careful consideration in order to provide a complete range with a minimum number of units. The sizes of industrial ignitron tubes for welding control and for rectifier service have been chosen on a preferred number scale providing a proportionate increase at each increment. The development of a 400-ampere ignitron for rectifier service was undertaken to provide the next larger size tube in the series. Development work on the 400-ampere ignitron disclosed that design problems become more complex as the physical size of the tube is increased. The ability of a tube to rectify a given current with a minimum number of arc-backs is affected by circuit duty and by control of ionization, particularly at the end of the current commutating period. None of the factors involved have indicated an upper limit in size, but rather a need for a more exact understanding of the phenomena involved.

Figure 1. Sealed ignitrons for welder service ranging in capacity from 300 to 2,400 kva at 250 to 600 volts (rms)

THE DEVELOPMENT of the ignitor introduced a new and different concept in the utilization of mercury pool cathodes for mercury arc rectifying devices.[1] The ignitor permitted the starting of a new cathode spot for each conducting period at the precise instant that conduction was desired. With this mechanism it was possible to design mercury pool tubes of the single-anode half-wave type having the control characteristics of the thyratron, the versatility of the half-wave tube in circuit application work, and the very high emission capacity of the mercury pool. This emission capacity seems partly caused by the ability of the cathode spot to divide and subdivide above currents of the order of 15 to 20 amperes until a sufficient number exist to supply the current demand of the circuit in which the rectifying element is connected.

Further recognition of the greater efficiency of the single-anode half-wave design for power conversion work in comparison with the older multianode rectifiers has been made recently in the development of an excitron tube.[2] With this tube, continuous excitation of a cathode spot on the mercury pool is maintained, and control of the starting of the anode current is accomplished electrostatically by means of grids as in the thyratron.

It is the purpose of this paper to review, briefly, the progress that has been made in the development and application of ignitron tubes and to outline some of the design considerations in the development of higher current sealed tubes.

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Status of Ignitron Tube Development and Application

Pumped Ignitrons

The largest field of application for continuously evacuated ignitron tanks has been in the electrolytic refinement of aluminum and magnesium. Here, large amounts of direct current are required for a single line of reduction cells. Capacities as high as 30,000 kw, 50,000 amperes at 600 volts are needed. Initially, to supply this capacity, individual rectifiers of the order of 3,000 kw were operated in parallel. The rectifying element of each unit consisted of 12 tanks having a continuous average current capacity slightly over 400 amperes per tank. Continued development supplemented by field experience has increased this capacity so that tanks of 1,000 amperes are practicable.

Welding Control Ignitrons

Initial application of permanently evacuated sealed ignitron tubes was made in an entirely new field—that of control and timing for resistance welding.[3]

In resistance welding, two pieces of metal are held together in close contact and a high current passed for a sufficient time to heat the parts to a welding temperature. The parts then are forged together. Control of the time is essential if uniform repetitive welds are to be made. Some progress had been made previously with the use of thyratrons to control the number of cycles in the weld, the current capacity of the thyratron being matched against the welder transformer primary requirements by means of a series transformer. Two ignitron tubes connected in parallel, but with current flow in opposite directions, provide a simple low-inertia switch with sufficient current carrying capacity to conduct the welding transformer primary current directly. In conjunction with thyratrons to initiate conduction, this precise control not only improves the quality of welds in materials

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fications of design to incorporate new factors were made until the one-half-per-cent tolerance was satisfied.

Conclusion

It is expected that the present a-c network analyzer will be expanded in the future to include sufficient units to permit solving almost any problems that may arise among the utilities of the area.

The objective of lowered costs has been achieved to some extent. Because of the research involved, its inseparability in the costs, the lack of production facilities, and the effect of wartime conditions on choice of materials and on labor obtainable, exact cost figures from the project are meaningless. However, it is believed that the use of electronic equipment and techniques has reduced the expense involved in erection of an analyzer by a factor of 40 to 50 per cent.

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usually fabricated by resistance welding, but also permits materials to be fabricated which heretofore had been considered too difficult to weld. As a result, welding methods have been extended from simple spot welds to line welding (a series of overlapping spots) to sequence control (in which provision is made for preheating, welding, forging, cooling, and annealing periods) and to flash welding. Sealed ignitron tubes developed for this service range in capacity from 300- to 2,400-kva demand for two tubes at supply voltages of 250 to 600 volts (rms). Figure 1 shows the general appearance of the welder ignitrons. All are water-cooled directly except the smallest, in which the heat is transferred to a heavy clamp that in turn may be either air- or water-cooled.

An interesting application of the ignitron contactor principle has been made to provide rapid interruption of the plate power supply in case of arc-over or short circuit in radio transmitters.[4] Ignitron contactors have been placed in each of the three 2,300-volt lines supplying power to the rectifier transformer. Ignition power is supplied through thyratrons. In case of fault, overcurrent operates small high speed d-c relays in the transmitting tube circuits that block the thyratrons, and the ignitron tubes in turn cease conduction. Interruption of fault currents occurs in slightly over one cycle of the 60-cycle supply frequency. Damage to the transmitter equipment from flashover and to the transmitter tubes is negligible.

Rectifier Ignitrons

Rectifier duty is more severe than welder control and it was necessary to modify the design of the sealed tubes somewhat, as discussed later, in order to provide service essentially free from arc-back. Two sizes of tubes were developed which in 300-volt d-c service have continuous average current ratings per tube of 100 and 200 amperes. At 600 volts, the current capacities are 75 and 150 amperes. In addition, overload capacity is required since rectifier equipment for industrial, mine, and railway service has standard 2-hour and 1-minute ratings. These standards vary, depending on the service, from 125 to 150 per cent at the 2-hour point and from 200 to 300 per cent at one minute. To meet these requirements, the continuous or design capacity of the tubes must approach the overload requirements because the mass of the tube electrodes is small, in general, and there is relatively low heat storage capacity.

With these two sizes of tubes, it was possible to provide sealed ignitron rectifier equipment ranging in capacity from 75 to 500 kw at 300 volts, and from 100 to 1,000 kw at 600 volts by proper choice of the circuit and number of tubes.[5] Trial installations were made in the industrial, mining, and electrolytic field, and for building service d-c supply. Sufficient experience had been obtained at the beginning of World War II to provide field tested and proved equipment for the great industrial expansion that took place.

400-Ampere Sealed Ignitron

In the initial development of steel envelope sealed ignitron tubes, there were two predominant considerations—vacuum life and ignitor life. Rectifier duty, in particular, is continuous in nature and it is essential that sealed tubes for this service have a long life to avoid the interruptions of frequent replacements. Experience has demonstrated amply that tubes properly degassed and tightly sealed retain the required vacuum conditions for years. In fact, the action of the arc and the walls of the tube exert a considerable cleanup action which tends to improve the initial vacuum.[6] Under ordinary service conditions, ignitors show very little wear or erosion. Under improper conditions, such as accidental reversal of the ignitor current, the life may be very short—a matter of minutes—since a cathode spot then forms on the ignitor and material is blasted away rapidly. Actual life of some of the tubes in initial installations has exceeded five years with these tubes still in service.

With substantiation through field experience of the basic factors of vacuum and ignitor life and the excellent performance of the higher current rectifiers for aluminum and magnesium reduction, it was felt that development of the next larger size of sealed ignitron tube was desirable and could be undertaken.

Size

It is fully as important to choose a proper series of sizes in electronic tubes as in other electric or mechanical devices. Preferred number scales which provide a proportionate increase in size or capacity at each increment offer a logical means of determining the number of sizes required to serve a given field. Sizes for both the welder and rectifier ignitron tubes were established on this basis. In the case of the rectifier ignitrons, the series 25, 50, 100, 200, 400, 800, 1250, 2500 gives a satisfactory spacing in average current per tube. At the lower end of the scale, 25 amperes and below, thermionic-cathode tubes of the thyratron type are used because of the reduction in required auxiliary equipment. The number of sizes required above the 400-ampere point will depend primarily on application trends. In general, it is preferable to use larger sized tubes rather than increased numbers for increased capacity, provided proper wave shape and induction co-ordination factors are maintained.

Design

Ignitron tubes are basically heat-transfer devices in which maximum temperatures establish the limiting operating conditions as in other forms of electric equipment. The limiting temperatures are not those, in general, at which gradual deterioration or burning takes place, but rather those at which complete failure of the rectifying properties occur. Such failures rarely cause destruction of the rectifying element itself unless the arc-backs are allowed to persist. Normally, restoration of conditions below the limiting ones restores the rectifying action.

The insulating patch theory developed by K. H. Kingdon and E. J. Lawton seems to offer the most reasonable and consistent explanation of the cause of arc-back.[7] Briefly, it is assumed that small microscopic insulating particles exist on the surface of the anode or are carried there by the vapor stream. These patches become charged from the ionization present. If the charge becomes sufficiently great and the distance separating the charge from the anode is sufficiently small, field emission will release enough electrons to establish a glow discharge. The resulting increase in ionization favors the establishment of a cathode spot on the anode, which is followed by arc conduction of current in the reverse direction. The probability of exactly the right conditions occurring depends on the density of ionization in the immediate vicinity of the anode, the rate of ion collection, the frequency of trial, and the size and leakage resistance of the insulating patch. For a given design, the majority of the factors depend on or are related to circuit requirements so that a knowledge of these provides a basis for predicting tube performance.

From the tube design view point, it is difficult to isolate the effects of the factors influencing arc-back since they are interrelated and interdependent. The more important ones are as follows:

1. Vapor pressure.
2. Ionization.
3. Anode current.
4. Anode voltage.
5. Circuit duty.

Vapor Pressure

Energy losses occur in an ignitron tube as a series of pulses. These pulsations are averaged by the mass of the enclosing envelope to an even flow of heat which is removed by the water-cooling. The water temperature therefore determines a base or datum plane above which the instantaneous pressure rises during the conducting cycle. The instantaneous pressures are the result of vaporization of mercury by the action of the cathode spot. Mercury is vaporized at a rate of approximately 7.2×10^{-3} grams per coulomb, and this mercury blast creates pressures during the conducting period far in excess of the pressure determined by the wall temperature. The pressure during conduction is of little importance as regards arc-back, except as its magnitude influences the ability of the condensing walls to reduce the pressure at the end of conduction to a value below the breakdown point. A typical pressure breakdown curve for static conditions in mercury is shown in Figure 2. Since the pressure waves travel with a velocity approximately that of sound, there is a time lag between the current and the corresponding instantaneous pressure at the end of conduction which tends to delay deionization. This lag increases as the size of the tube increases.

In addition to the vaporized mercury, the action of the cathode spot ejects relatively large (visible) mercury droplets which travel at a rate very much slower than the pressure wave. The droplets may be observed in glass tubes by blocking out the glow during conduction with a synchronously rotating slotted disk and by illuminating the vacuum space with the light from a tungsten filament lamp. The initial velocities of individual drops vary widely since droplets are visible at the top of their trajectories over nearly the entire length of the tube. These droplets undoubtedly create high transient pressures when they strike relatively high temperature objects, and because of their random character may account for some of the arc-backs that are found to occur late in the inverse cycle when both the residual ionization and inverse voltage have decreased appreciably from their maximums.

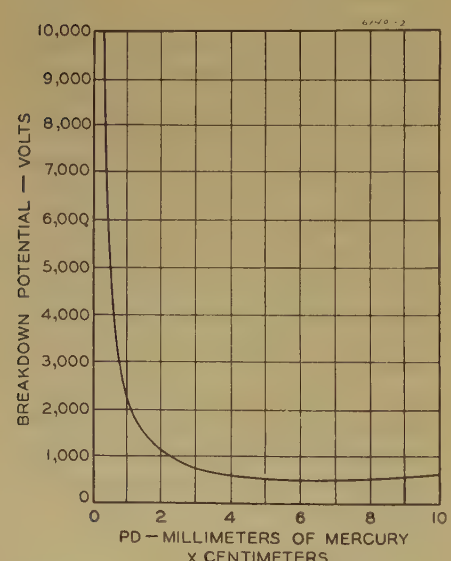
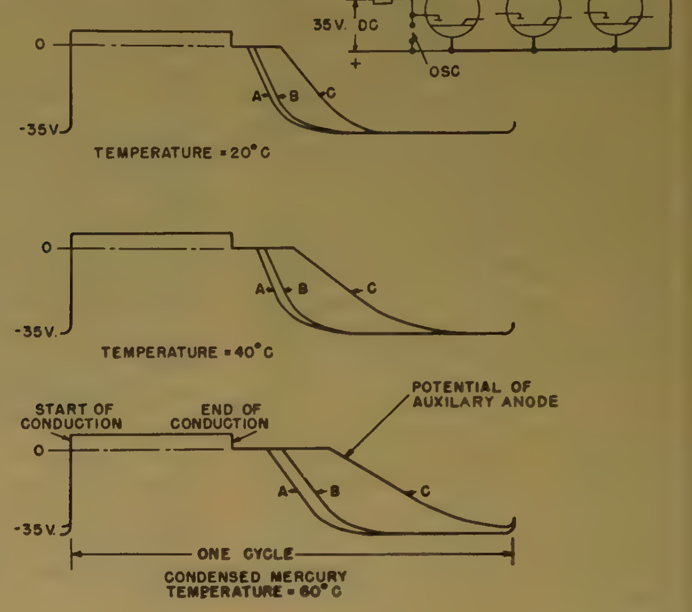


Figure 2. Breakdown voltage between two electrodes in mercury vapor as a function of the product (Paschen's law) of the vapor pressure *P* and the gap length *D*

Figure 3. Comparative deionization rates in 100-, 200-, and 400-ampere tubes as indicated by the potential of a probe

A—$3^1/_2$ inch diameter tube
B—5 inch diameter tube
C—8 inch diameter tube



Ionization

The problem of ionization becomes increasingly important as the size and capacity of the tube is increased.

Under low pressure conditions, the mean free path of the ions is relatively long, and the arc voltage needed to produce the required ionization is relatively high. Under extremely low pressures, there is a condition in tubes having constricted arc paths where there are insufficient ions to neutralize the space charge and for an instant the tube behaves as though it were a high vacuum tube.[8] The current is reduced abruptly, and this change in conjunction with associated circuit inductance produces high voltage surges. The point at which pressures become low depends largely on the tube design. In tubes of small volume and open construction, the possibility of constriction surges is remote at any temperature above zero degree centigrade (vapor pressure, 0.16 micron) because the action of the cathode spot very rapidly establishes the vapor density required to provide the necessary ions. As the size of the tube increases, and particularly as grids are added to control the ionization, the minimum pressure must be increased. In multiple grid high voltage tubes, minimum pressures of the order of four to eight microns (35–45 degrees centigrade) are required for currents of 4,000 amperes. The minimum pressures usually are determined by the short-circuit current that the tube may be required to conduct, rather than the normal load currents.

At the end of conduction, the intensity of residual ionization becomes extremely important because of its direct relation to arc-back. In tubes of small volume, the deionization rate is rapid and equilibrium between ionization and anode current is reached quickly so that the ion density is relatively low when inverse voltage is applied at the end of commutation, or transfer of current from anode to anode. As the cross section and volume of the tube increase, for higher capacities, the volume of ionized gas increases and the time required for the ions to diffuse to the walls is increased, with the result that the density of ionization at the end of conduction is correspondingly higher. Figure 3 shows the time required for the ionization to decay to a given value for a fixed value of anode current in the 100-, 200-, and 400-ampere sizes of tubes. The curves repre-

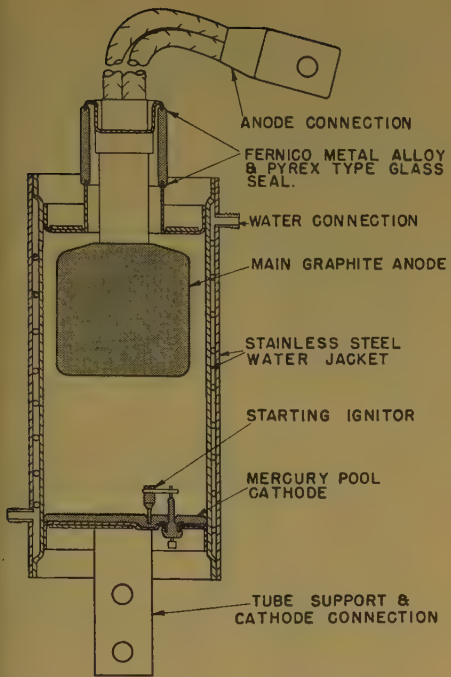


Figure 4. Cross section of a 2,400-kva control ignitron

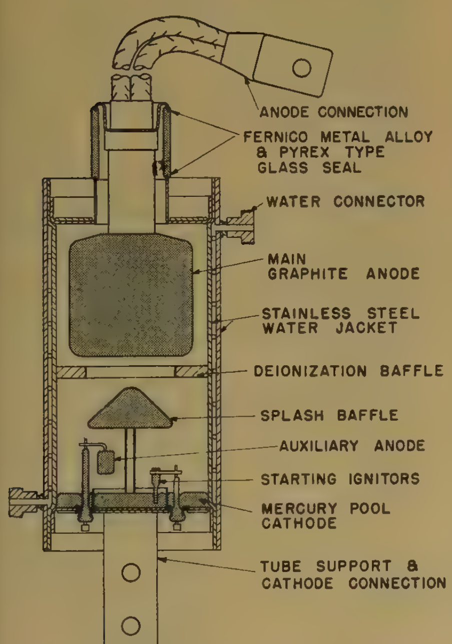


Figure 5. Cross section of a 200-ampere rectifier ignitron

sent the auxiliary anode-to-cathode voltage as seen on a cathode ray oscilloscope, with the auxiliary anode used as a probe and connected to a negative bias to collect positive ions as indicated in the connection diagram insert. The point in the curves at which the auxiliary anode voltage begins to increase negatively indicates the time required for the ionization to decay to a value just at which the probe current can be supplied without an increase in voltage. The effects of increased tube diameter, arc volumes, and particularly increased vapor density are clearly shown.

In the development of the 400-ampere tube, it was expected that the general design of the smaller tubes could be extended to this size. However, operation tests indicated that both the ionization density and time of deionization must be decreased. One of the most effective methods of accomplishing this in the larger tubes is to enclose the anode within a grid. The grid functions primarily as a shield to isolate the anode from the ionization in the arc chamber. The small volume of ionization between the grid and anode insures rapid deionization of this space. Probably of equal importance is the current density level that can be collected on any part of the anode.

Anode Current

The magnitude and wave shape of the anode current influences the magnitude and wave shape of the vapor pressure wave and affects the ionization density. These factors depend, in general, on the mode of operation of the rectifier circuit employed. The delta 6-phase double-Y circuit (proposed AIEE Standards nomenclature) commonly is used and in this circuit there is a ratio of three to one between the peak and average currents. This ratio usually varies from two, three, four, six to one, depending on the type of circuit. At the end of conduction, the current wave shape, and particularly the magnitude of the ionization resulting from the current, depends on both the commutating reactance of the transformer and supply and the voltage that is available to transfer the current from one phase to another.

Anode Voltage

The degree of positive ion bombardment and the thickness of the ion sheathes that form around the anode during the negative part of the cycle are functions of the magnitude of the inverse anode voltage. This voltage therefore influences the factors involved in arc-back directly. The magnitude of the inverse voltage depends on the circuit and is higher for circuits having low ratios of peak-to-average currents and lower for those having high ratios. Two periods in the inverse voltage wave are of particular interest. The initial inverse voltage occurs at the end of commutation at the time the residual ionization is highest. Its magnitude depends on the circuit, the transformer commutating reactance, the current commutated, and the amount of phase control. The maximum inverse voltage depends on the circuit and usually occurs later in the

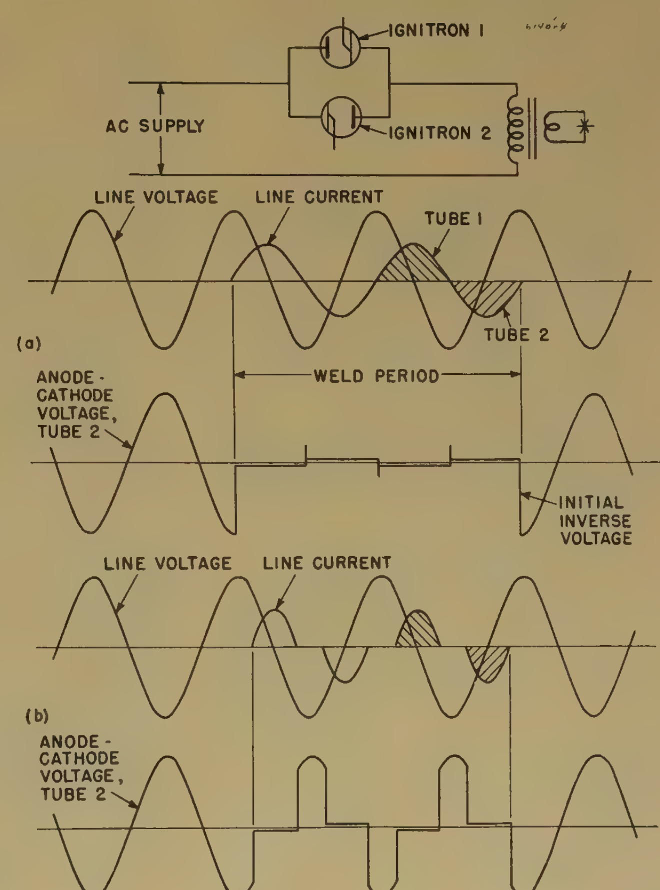


Figure 6. Tube voltage and current wave shapes for welding control operation

(a). No phase retard
(b). Approximately 30 degrees phase retard

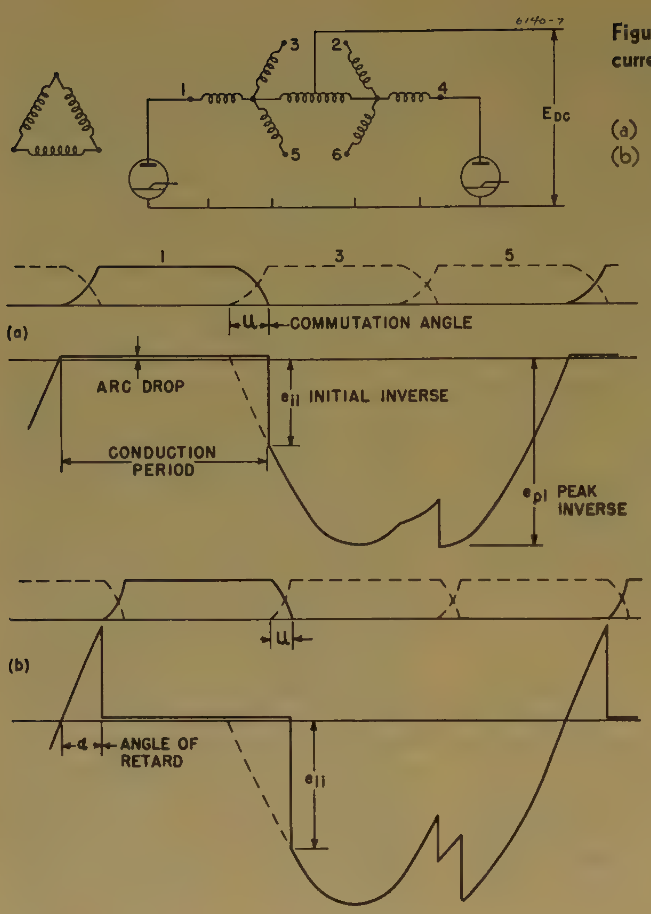


Figure 7. Tube voltage and current wave shapes for rectifier operation

(a) No phase retard

(b) Approximately 30 degrees phase retard

cycle when the residual ionization is reduced. Under large amounts of phase control (70-to 100-per-cent voltage reduction), the initial inverse voltage approaches or equals the maximum in magnitude.

Circuit Duty

The influence of the circuit duty on the tube design is shown in Figures 4 and 5. The figures show, respectively, a sealed ignitron for welder service and one for rectifier service. The two tubes have essentially the same physical dimensions and operate at inverse voltages of about the same magnitude.

In welder service, two tubes are connected back to back to conduct both halves of the a-c wave. The conditions

Table I. Tube Ratings

	Maximum Anode Current, Amperes: 300 Volts Direct Current	Maximum Anode Current, Amperes: 600 Volts Direct Current
Instantaneous	3,600	2,400
Average		
Continuous	400	300
2-hour	600	450
1-minute	800	600
Surge (0.15 second maximum)	25,000	19,000

for welder service are shown in Figure 6 for several current cycles in a single spot weld, followed by an off period before the next weld. For the no-phase retard conditions shown in Figure 6a, only the last tube to conduct in the final cycle must withstand an inverse voltage when appreciable ionization is present. Under phase controlled conditions as shown in Figure 6b, the current wave becomes discontinuous so that each tube must withstand commutating conditions in each conducting cycle, and the resulting duty therefore is increased by the increased number of trials. Recognition of this is made in the recommended application practice by reducing the peak current in proportion to the phase control.

In rectifier service, the tubes must withstand successfully commutating conditions in every cycle. Figure 7 shows the tube conditions in a delta 6-phase double-Y rectifier with no-phase retard (a) and phase retard (b). The tube current at commutation is a segment of a short-circuit current wave whose magnitude is limited primarily by the commutating impedance of transformer and by the tube arc-drop. Under phase retard conditions, the voltage that is available to force current transfer from one transformer winding to another is increased, with the result that the transfer time of a current of given magnitude is decreased. The residual ionization at the instant inverse voltage is applied is therefore higher, and this in combination with the higher inverse voltage increases the duty.

In the foregoing discussion of design problems, it should be emphasized that the problem of arc-back has been considered at limiting conditions and that a reasonable reduction in operating conditions below the limit results in performance practically free from arc-back.

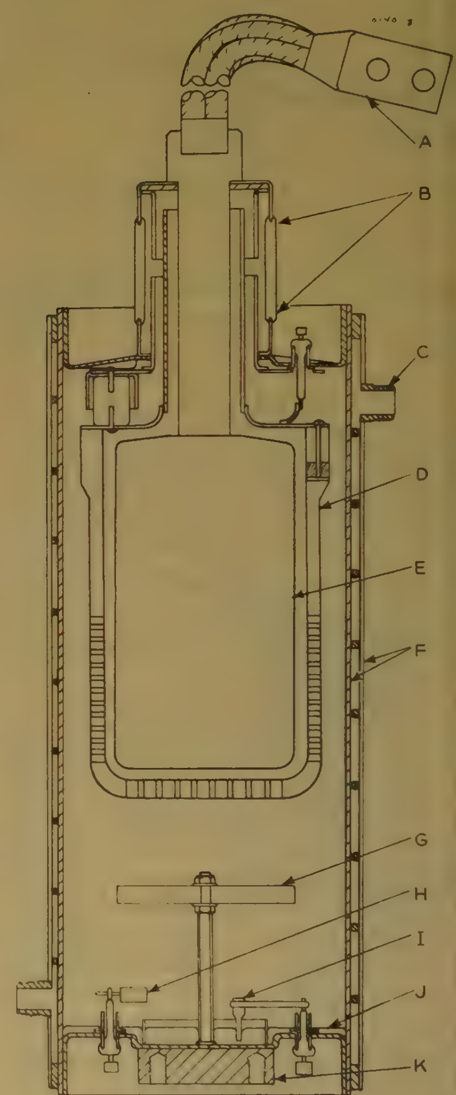


Figure 8. Cross section of the 400-ampere ignitron showing the general design

A—Anode connection
B—Fernico metal alloy and Pyrex type glass seal
C—Water connection
D—Graphite grid
E—Main graphite anode
F—Stainless steel water jacket
G—Splash baffle
H—Auxiliary anode
I—Starting ignitors
J—Mercury pool cathode
K—Tube support and cathode connection

Table II. Combinations of Tubes and Circuits

Kilowatt Output 300 Volts	Kilowatt Output 600 Volts	Number of Tubes	Circuit
400	750	6	Delta 6-phase double-Y
500	1,000	6	Delta 6-phase double-Y
750	1,500	12	Delta, 12-phase quadruple zig-zag Y
1,000	2,000	12	Delta 12-phase quadruple zig-zag Y

Construction

The general design and construction of the 400-ampere tube is shown in Figure 8. The essential materials and general design of the earlier ignitron tubes have been retained. The water-cooling jackets and the inner cylinder forming the vacuum chamber are of stainless steel to minimize corrosion. The anode insulating and supporting bushing is of the fernico-glass sealing combination that has proved very satisfactory, not only from the vacuum tightness standpoint but from the viewpoint of strength and shock resistance as well. Anode and grid are of special graphite. Two ignitors are incorporated again to provide for circuit reversal or other accidents. Only one is required for ignition. An auxiliary anode provides cathode spot stability for anode currents below three to five amperes peak, and permits simplification of the ignition circuit control.

A completed tube is show in Figure 9. The water jacket is approximately nine inches in diameter and 22 inches tall. Over-all height, including the flexible anode lead, is 40 inches, and the tube weight is approximately 95 pounds.

Figure 9. A completed 400-ampere sealed ignitron

In the larger sizes of tubes, the materials used in the construction become an increasing proportion of the tube cost, and, even though warranties protect against loss from early failure, it becomes desirable to provide in the design the possibility of repair. In this tube the header closing welds are of the fusion type rather than resistance seam welds. This permits removal of the headers with relative ease for inspection or repair.

Performance and Rating

Essential ratings of the 400-ampere tube are given in Table I, and Table II shows possible combinations of tubes and circuits to provide industrial rectifiers of standard capacities at the 300- and 600-volt levels.

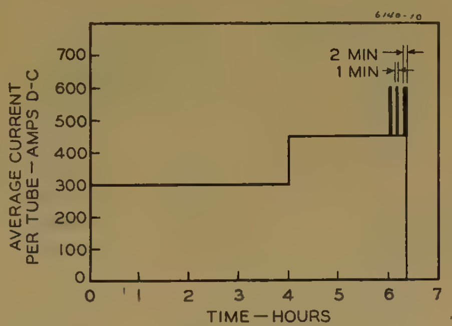


Figure 10. A 600-volt operation test at rated continuous 2-hour and 1-minute load currents for zero phase retard

Rectifier performance tests usually are made on the load increment plan for a chosen water temperature and flow and at the desired operating voltage levels.[9] The commutating impedances are adjusted to give output voltage regulation and short-circuit currents comparable to these expected in actual service. The load usually is absorbed by motor generator sets.

Prior to determining limits, an operation run usually is made to check the performance at rated continuous 2-hour and 1-minute loads. Figure 10 shows the time-current plot of such a run for operation at 600 volts direct current.

Figure 11 shows the general shape of the load limit current curve for the 400-ampere ignitron as the output voltage of the rectifier is reduced progressively to zero by means of phase control. Limits as far as the tubes are concerned are determined by arc-back. Under phase retard conditions, the combined effects of a higher initial inverse voltage and greater ion density produce a higher tube duty, which is reflected by a reduction in load current.

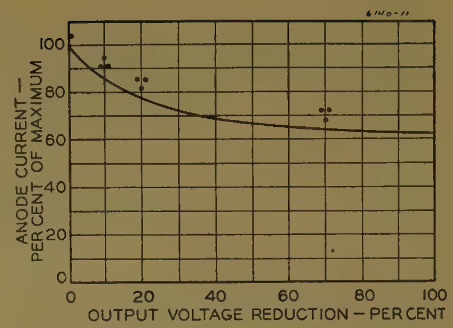


Figure 11. Continuous average anode current, determined by load limit tests, as a function of output voltage reduction by phase control

The dots indicate arc-backs

Conclusion

The development of the 400-ampere sealed ignitron revealed no factors which would tend to limit the size and capacity that could be designed. Undoubtedly, larger tubes would require different treatment and probably different forms. The development did show the very great part played by ionization and the need, in spite of the excellent fundamental work that has been done, for additional work with particular emphasis on the problem of rectification.

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The Application of Series Capacitors to Flash Welders

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ASSOCIATE AIEE

THE APPLICATION of series capacitors to spot, projection, and seam welders no longer presents any difficult problems, and they have been used successfully with these types of welders for several years.

Essentially, enough capacitive reactance is placed in the primary circuit of the welder transformer to neutralize the inductive reactance of the welder. This theoretically produces unity power factor and reduces the kilovolt-ampere demand to the level of the kilowatt demand. Since the voltage across the capacitor units is directly proportional to the current flowing through them, units must be selected with a voltage rating high enough so that this rating is not exceeded unless they are protected by an overvoltage device.

The flash welder, however, presents an entirely different problem since the welding operation is composed of two phases which differ widely in both power factor and kilovolt-ampere demand.

The first phase is termed flashing and occurs when the pieces to be welded are brought in contact and heated to a plastic state by the passage of the current as it flows from one piece to the other. There is some doubt as to whether this current flow is caused by metallic or gaseous conduction. However, we do know that there is not good contact between the two pieces, and we will speak of the condition during current passage as an arc. The flashing kilovolt-ampere demand is less than the final upset phase, the power factor is relatively high, and the wave shape of the current is rather irregular, as shown in Figure 1.

When the plastic temperature has been reached, the upset phase takes place. Here the pieces are forced together quickly and are forged, producing the complete weld. Since the resistance at the weld region is lower at upset than when flashing, the kilovolt-ampere demand is greater, the power factor lower, and because of the absence of arcing, the wave shape of the current is smooth, as illustrated in Figure 2.

An example of this type of machine is the barrel welder. Figure 3 is a simplified drawing of the essential parts. Note that as the cam appearing at the extreme right revolves, the movable platen, in which is clamped one edge of the sheet forming the barrel, moves slowly toward the edge of the sheet which is clamped in the stationary platen. The edges are brought in contact as the cam begins rotating, and, because of the inherent irregularities and minute projections along the edges of the stock, the contact resistance is high and the current flow is restricted to small areas. This produces a flashing action which burns away the multiplicity of projections. It is evident that there are two current paths in parallel: one short path through the weld region and the other, a longer path, through the cylinder. The relatively high resistance offered by these two paths in parallel therefore results in a comparatively low kilovolt-ampere demand and high power factor.

When the upset block on the cam meets the movable platen, the edges of the stock are forced together, the arc extinguished, and, because of the resulting decreased resistance, the current flowing through the weld region increases greatly. The ratio of the resistance to the inductive reactance of the secondary circuit decreases with an accompanying decrease of power factor.

A series of tests were made on a standard 350-kva 440-volt 42-inch barrel welder welding a standard 16-gauge 55-gallon drum to determine the possibility of applying series capacitors to reduce the kilovolt-ampere demand and increase the power factor.

Oscillograms were taken before any power factor correction was attempted and it was found that the demand during flashing was about 350 kva at 92 per cent power factor, and at upset the demand increased to 1,200 kva at a power factor of 48 per cent. The relative magnitudes are shown graphically in Figure 4.

Suppose we add enough capacitance to

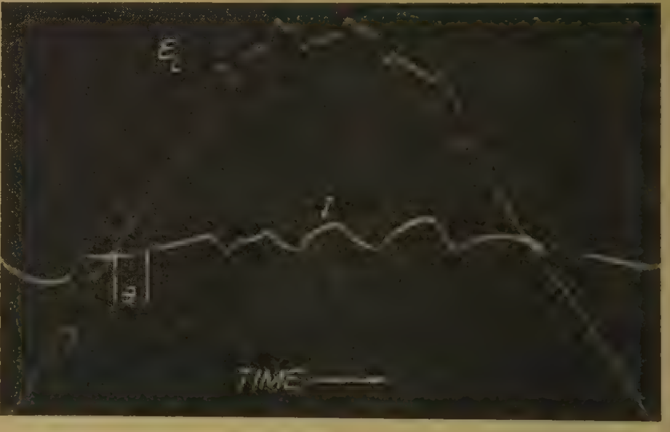

Figure 1. Oscillogram of current and voltage during flashing

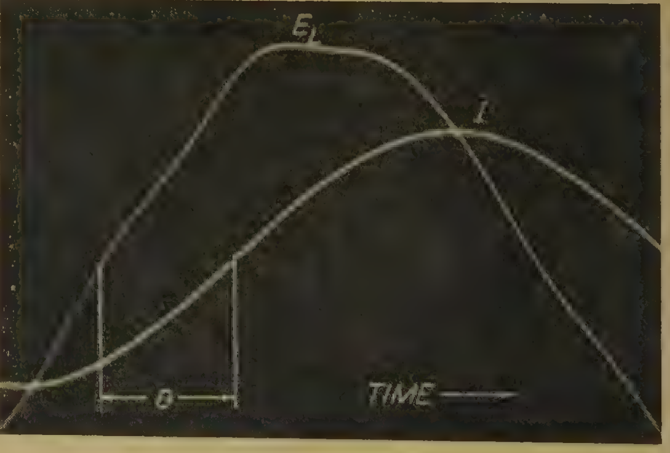

Figure 2. Oscillogram of current and voltage during upset

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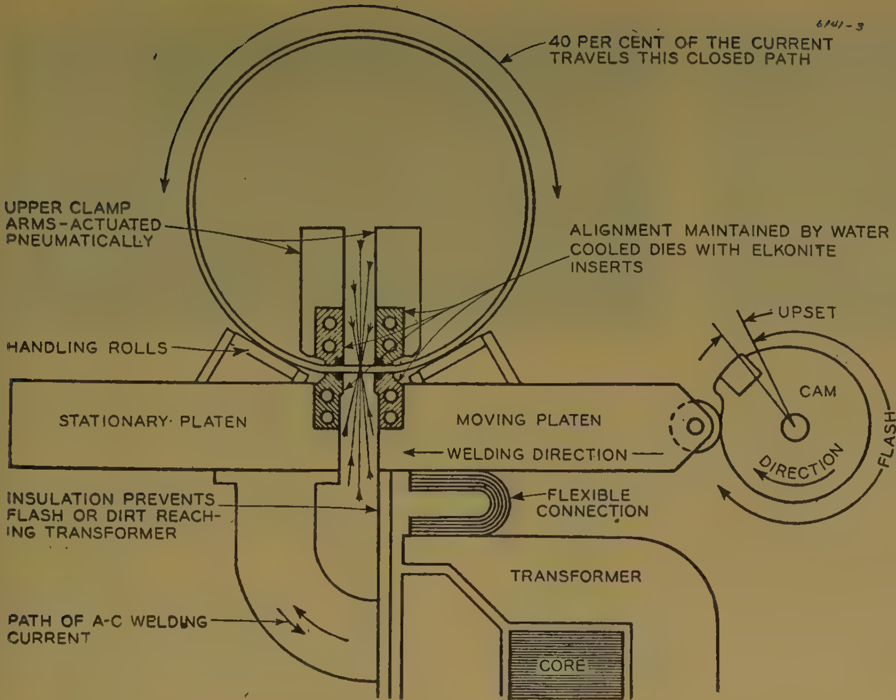


Figure 3. Essential parts of a drum welder

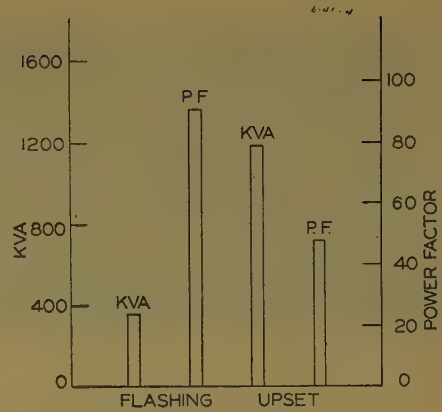


Figure 4. Initial uncorrected values of kilovolt-ampere demands and power factors

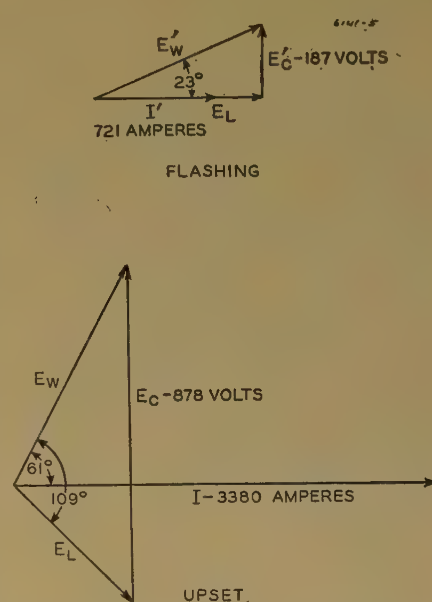


Figure 5. Vector diagrams of currents and voltages existing when the flashing period is corrected to unity power factor

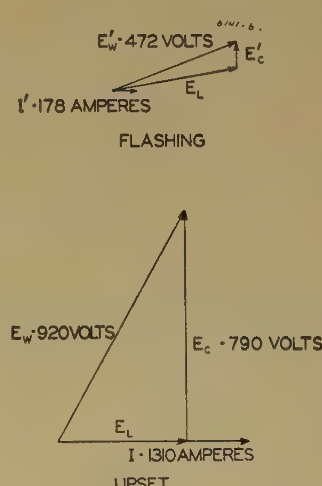


Figure 6. Vector diagrams of currents and voltages existing when the upset period is corrected to unity power factor

the primary circuit of the welder to obtain approximately unity power factor during the flashing period. Proceeding exactly as is done in applying series capacitors to any other type of welder, it is found that 29 230-volt capacitors are required. Figure 5 gives vectorially the voltages existing in the circuit at this time.

During the upset stage of the weld, the current increases, thereby producing a proportional increase in the primary current which now flows through the capacitor bank. This results in a voltage drop across the capacitors which is approximately 150 per cent greater than the maximum permissible voltage of the units. Of course, units with a greatly increased voltage rating could have been chosen initially, but the number required would have been approximately four times that actually selected, so that this is clearly not an economical solution. The demand also has increased to a value approximately 25 per cent greater than it was before the capacitors were applied. The upset portion of Figure 5 gives these values.

Let us now increase the capacitance to that required to correct the upset portion of the weld to unity power factor. We find that it is necessary to use 36 575-volt units. However, the resulting series circuit at upset impresses a voltage of about $2^1/_2$ times the normal supply voltage across the welder transformer, as shown in Figure 6. There are two methods that can be used to maintain the secondary voltage appearing across the platens at the value it was before the capacitors were applied. One method is to increase the welder transformer ratio, and the other is to use a step-down transformer in the supply circuit. In this instance let us choose the first method and increase the transformer ratio.

As shown in the flashing section of Figure 6, we immediately meet complications during the flashing period since the characteristics of the circuit now are such that the voltage impressed on the welder transformer primary is reduced to 60 per cent of that required for proper flashing. This would result in an imperfect weld.

We now are able to conclude that, when there is an appreciable difference between the flashing and upset demands, we will have one of two undesirable conditions present if we apply series capacitors to flash welders, according to conventional theory:

1. Correction of the flashing demand will result in excessive condenser voltage during upset.

2. Correction of the upset demand will result in a welder transformer voltage lower than that required for proper flashing.

Since, in general, it is not economical to use high voltage capacitor units, we will provide a means of increasing the voltage across the welder transformer during the flashing period and use a value of capacitance which will give unity power factor during upset, since

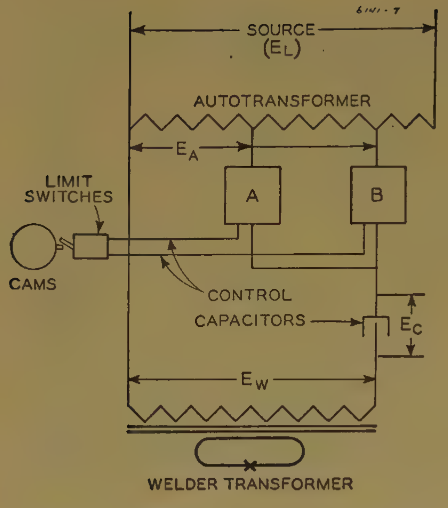


Figure 7. Schematic diagram of circuit used in test

the power factor at flashing is rather high without correction.

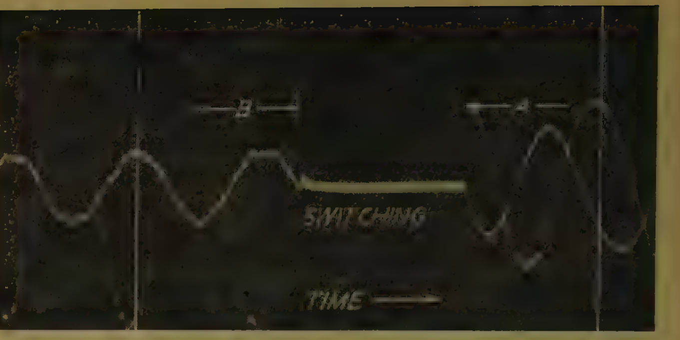
In order to accomplish this condition, the circuit shown in Figure 7 was made. A cam has been placed on the shaft which rotates the platen-moving cam, a limit switch which is actuated by this cam has been added, a second tube contactor has been placed in parallel with the original contactor, and an autotransformer has been connected in the supply line. The series capacitors are placed in the circuit between the contactors and the welder transformer primary winding.

The sequence of operation of the circuit is as follows.

When the limit switch is closed, contactor *B* is closed and contactor *A* is left open. The tap on the autotransformer to which contactor *B* is connected is selected so that approximately normal supply voltage is impressed on the welder transformer. This energizes the circuit correctly for the flashing operation. The added cam is adjusted so that contactor *A* is closed and contactor *B* is opened about one cycle before the platen-moving cam starts the upset movement. This adjustment was necessary as there was an interval of one cycle during the switching of contactors in which there was no current flow. However, it is possible to eliminate this period by the use of proper controls. Figure 8 shows this interval. Contactor *A* is connected to a lower voltage tap on the autotransformer and results in normal supply voltage being impressed on the welder transformer during upset.

In order to simplify the calculations, the assumption is made that the source is of infinite capacity. This assumption changes the values slightly from those actually existing during the tests.

Figure 8. Oscillogram of contactor switching interval



The first step is the calculation of the inductive reactance of the welder during upset. This calculation is the value, in ohms, of the capacitive reactance that must be supplied by means of the capacitor units in order to obtain approximately unity power factor at upset. The circuit voltage conditions are as shown by the vector diagram in solid lines in Figure 9. It is evident that the voltage to be supplied by the autotransformer is the welder voltage E_W multiplied by the power factor during upset, plus the tube contactor drop of approximately 15 volts.

The next step is the determination of the circuit constants during flashing. The net reactance is the difference of the capacitive and inductive reactances. This value then is combined vectorially with the flashing resistance to find the resulting impedance. The voltage at the autotransformer then must be equal to this impedance multiplied by the primary flashing current. This current is approximately equal to that occurring during the uncorrected flashing condition. See Appendix II for exact calculation.

The vector diagram composed of broken lines illustrates the voltages existing during flashing. The power factor is slightly lagging because of the fact that the inductive reactance during flashing is greater than during upset. We are not certain as to the exact reason for this change in reactance, but it may be caused by the harmonics present during flashing.

Figures 10 and 11 illustrate oscillograms of the current and voltage as measured at the supply side of the autotransformer after the capacitors had been applied. Note that the power factor during upset is not quite unity and that the power factor during flashing is less than at upset.

Figure 12 shows graphically a comparison between the kilovolt-ampere demand and power factors before and after applying correction in the actual test, the dashed lines giving the corrected values.

It is evident from an examination of Figure 12 that while the low kilovolt-ampere demand and high power factor of the flashing phase has not been altered appreciably, the upset kva demand has been reduced to approximately one-third of its uncorrected value, and the power factor has been increased from 48 to 95 per cent.

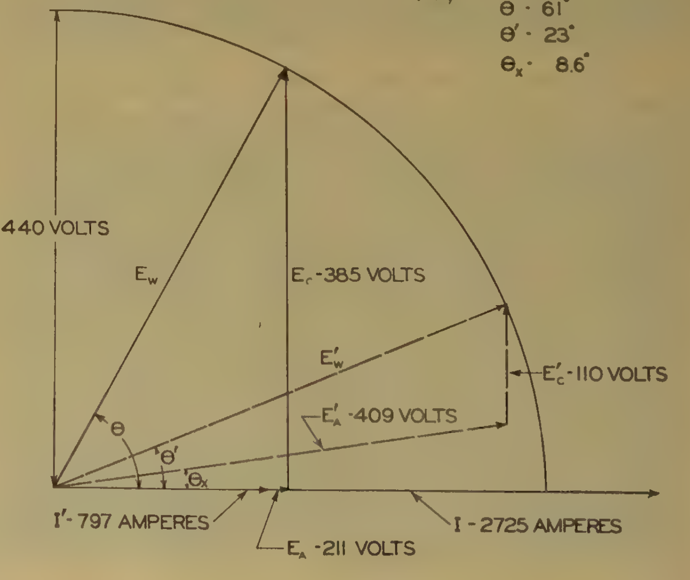


Figure 9. Vector diagram of currents and voltages existing during flashing and upset after capacitors have been applied in final arrangement

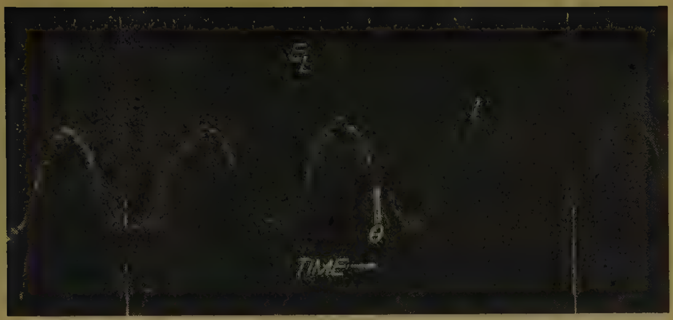


Figure 10. Oscillogram of current and voltage during flashing after addition of capacitors

Figure 11. Oscillogram of current and voltage during upset after addition of capacitors

Appendix I. Notation

E_L = supply line voltage
E_A = autotransformer output voltage
E_C = capacitor voltage
I = output current of autotransformer
θ = angular difference between current and voltage in high tension winding of welder transformer

The foregoing refer to upset values. With prime notation they refer to flashing values.

θ_x = phase difference between current and voltage in supply line during flashing after capacitors had been applied

Appendix II. Calculations

Upset period:

$$Z=\frac{E_L^2\times 10^{-3}}{\text{kva}}=\frac{440^2\times 10^{-3}}{1,200}=0.161 \text{ ohm}$$

$$X_L=Z\times \sin 61 \text{ degrees}$$
$$=0.161\times 0.875$$
$$=0.141 \text{ ohm}$$

Therefore $X_C=X_L=0.141$ ohm and power factor becomes theoretically unity.

$$E_A=E_L\times \cos 61 \text{ degrees}$$
$$=440\times 0.48$$
$$=211 \text{ volts}$$
$$E_C=E_W\times \sin 61 \text{ degrees}$$
$$=440\times 0.875$$
$$=385 \text{ volts}$$

Corrected kva = uncorrected kva × cos 61 degrees
$$=1,200\times 0.48$$
$$=575 \text{ kva}$$

$$I(\text{corrected})=\frac{\text{kva (corrected)}}{E_A\times 10^{-3}}$$
$$=\frac{575}{211\times 10^{-3}}$$
$$=2,725 \text{ amperes}$$

Flashing conditions:

$$Z'=\frac{E_L^2\times 10^{-3}}{\text{kva}}$$
$$=\frac{440^2\times 10^{-3}}{350}$$
$$=0.554 \text{ ohm}$$
$$X'_L=Z'\times \sin 23 \text{ degrees}$$
$$=0.554\times 0.39$$
$$=0.218 \text{ ohm}$$
$$X_L(\text{net})=X'_L-X_C$$
$$=0.218-0.141$$
$$=0.077 \text{ ohm}$$
$$R'=Z'\times \cos 23 \text{ degrees}$$
$$=0.554\times 0.92$$
$$=0.51 \text{ ohm}$$

Thus the power factor with capacitors is

$$\cos^{-1}\tan\frac{X_L(\text{net})}{R'}$$
$$\tan\theta_X=\frac{X_L(\text{net})}{R'}$$
$$=\frac{0.077}{0.51}$$
$$=0.151$$
$$\theta_X=8.6 \text{ degrees}$$
$$\cos\theta_X=0.988$$

Power factor = 98.8 per cent

$$E'_A=\frac{E'_W\times \cos 23 \text{ degrees}}{\cos 8.6 \text{ degrees}}$$
$$=\frac{440\times 0.92}{0.988}$$
$$=409 \text{ volts}$$
$$E'_C=E'_W\times \sin 23 \text{ degrees}-E'_A\times \sin 8.6 \text{ degrees}$$
$$=(440\cdot 0.39)-(409\cdot 0.1495)$$
$$=171-61.2$$
$$=109.8 \text{ volts}$$

Kilovolt-amperes (corrected)
$$=\frac{\text{kva}\times \cos 23 \text{ degrees}}{\cos 8.6 \text{ degrees}}$$
$$=\frac{350\times 0.92}{0.988}$$
$$=326 \text{ kva}$$

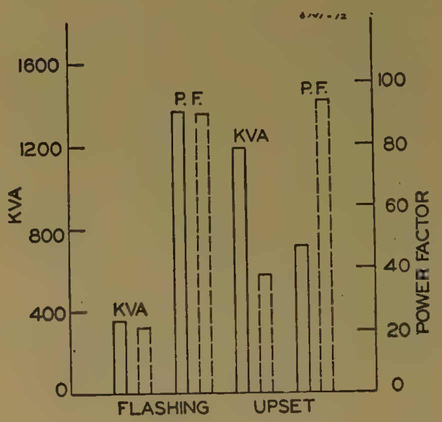


Figure 12. Kilovolt-ampere demands and power factors measured in test after addition of capacitors

Table I. Circuit Constants Before Correction

Circuit Constant	Value
Flashing	
Kva	350
Power factor	92 per cent
θ	23 degrees
E_L	440
Upset	
Kva	1,200
Power factor	48 per cent
θ	61 degrees
Supply voltage	440

$$I\text{ (corrected)}=\frac{\text{kva (corrected)}}{E'_A\times 10^{-3}}$$
$$=\frac{326}{409}\times 10^{-3}$$
$$=797 \text{ amperes}$$

The actual values of the autotransformer output voltages necessarily will have to be approximately 15 volts greater than the values as calculated to compensate for the voltage drop in the ignitron contactors.

Co-ordination of Insulation and Spacing of Transmission Line Conductors

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FELLOW AIEE

A SURVEY of practice on existing transmission systems shows, for the same operating voltage, wide variations in the insulation used. Also, wide variations occur in the spacing between conductors and from them to grounded members of the supporting structure.

These variations are illustrated in Figure 1, showing practice in the use of pin type insulators;[1] Figure 2, showing practice in the use of suspension insulators;[2,3,4] Figure 3, showing actual spacings between conductors for horizontally arranged lines;[4] and Figure 4, showing actual spacings between conductors for vertically arranged lines.[4,5] These variations possibly reflect to some extent the different degrees of lightning, wind, snow, and sleet encountered in different parts of the country, as well as the difference in experience and conservatism of the designers.

A recent committee report[4] gives data on 35 transmission lines in the 190- to 287.5-kv range. In this report is given a comparison of the impulse flashover of the insulators and the flashover from the conductor to the structure through air, with the conductor swung at 30 degrees from the vertical. In many cases the flashover to the structure through air is considerably less than the flashover of the insulators, indicating a lack of co-ordination in the design.

It is obvious from the preceding discussion that very little guidance can be obtained from existing lines as to the suitable number of insulators and spacings for projected transmission lines of various voltages. The writer, therefore, has undertaken to set up criteria for such insulation and spacings which will offer a guide for those engaged in this work, and which will be consistent for the various voltages.

In order to simplify the problem, the present study will apply mainly to steel tower lines with suspension insulators, although the question of wood pole lines and pin type insulator lines will be touched upon.

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Criteria for Minimum Line Insulation

As a basis for the selection of the minimum line insulation, the following criteria are suggested:

1. The 60-cycle wet flashover must equal or exceed the voltages expected from faults and switching surges. The maximum to be expected from these causes is five to six times normal line-to-neutral voltage. These voltages may be reduced by proper neutral grounding, by the use of breakers with minimum tendency for restriking, and by adequate lightning arrester protection. Under the most advantageous conditions, the upper limit of low frequency voltages may be taken as 3 to 3.5 times normal line-to-neutral voltage, and the 60 cycle wet insulator flashover should equal or exceed this value.

2. The impulse flashover should equal or exceed the standard basic impulse insulation level for the terminal equipment.[6]

In Table I are given the 60-cycle dry and wet and impulse flashover values of standard suspension insulators, ten inches in diameter and spaced $5^3/_4$ inches apart.[7]

In Table II are given 60-cycle and impulse flashover values for standard rod gaps.[7] These values may be used for gaps in air between conductors, between conductors and tower, and so forth.

In Table III, column 2, are given for the standard insulation classes the number of insulator disks spaced $5^3/_4$ inches apart, whose wet 60-cycle flashover will equal or exceed 3.5 times line-to-neutral voltage. This is the smallest number of insulators which will have an impulse flashover equal to or greater than the basic impulse insulation level.

Also given in Table III are the numbers of disks whose wet 60-cycle flashover will equal or exceed 4, 4.5, and 5 times the line-to-neutral voltage. In column 6, for comparison, are given the numbers of disks representing average practice, from Figure 2.

It will be noted that average practice follows column 5 up to and including 115 kv, column 4 for 138 to 230 kv inclusive, and column 3 for 287 and 345 kv. These

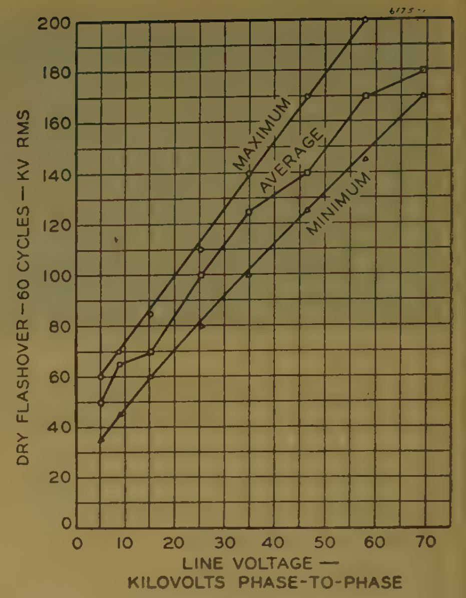


Figure 1. Practice in the use of pin type insulators

average values are from a curve. Actually, the only 287-kv lines in existence have the equivalent of 21 disks spaced $5^3/_4$ inches apart,[4] and so far there are no 345-kv lines. The tendency to use a greater relative insulation at the lower voltages is similar to the tendency to be found in the insulation of apparatus.

Co-ordination of Spacing With Insulation Level

With each level of insulation, there is a proper spacing between conductors and to the steel structure to develop the full strength of the insulator string selected. Let us take the minimum spacing to the steel structure *A* (Figures 5 and 6) as a spacing that will have an impulse flashover ten per cent greater than the impulse flashover of the insulator string.

Spacing *B* (Figures 5 and 6), the horizontal distance from conductor to tower when the conductor is hanging vertically, will be such that when the conductor is swung to 30 degrees from the vertical the impulse flashover from conductor to tower (spacing *A*) will be ten per cent greater than the insulator string flashover. It is assumed that during the summer season, when lightning prevails and there is no ice or snow, 30 degrees is ample to consider for the conductor swing.

Having determined spacing *B* from conductor to tower, spacing *D* between conductors for the horizontal arrangement (Figure 5) becomes two times spacing *B*, plus the width of the tower leg.

For the vertical arrangement (Figure 6), the spacing between horizontal planes through the conductors, spacing *C*, is a

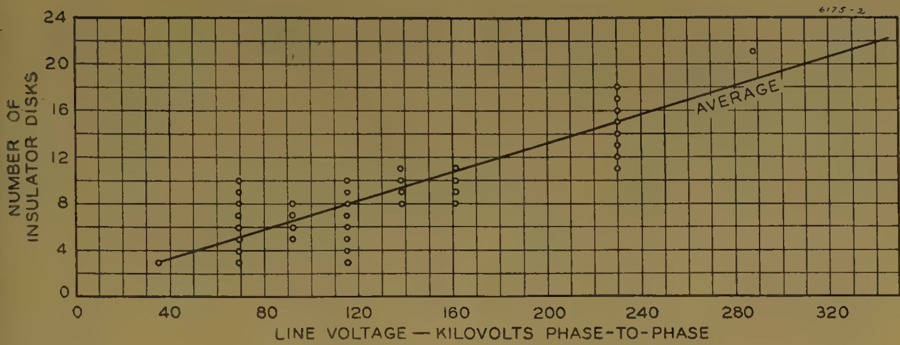


Figure 2. Practice in the use of suspension insulators

minimum of two times the spacing A, since the insulator string itself has a length approximately equal to spacing A. On the tower illustrated in Figure 6, on account of the slope of the crossarm bracing, the spacing between conductor planes naturally will be somewhat greater than $2A$. In other types of towers without crossarm bracing, this may not be the case, and the minimum spacing $2A$ can be used.

In order that the conductors may be fully shielded from direct strokes, it has been found that the two overhead ground wires are necessary, either for horizontal or vertical arrangement of conductors. If the two ground wires are placed close together, they blanket each other and the full benefit of the two wires is not secured. The effect is much the same as the effect of two rods driven in the ground. If the rods are close together, the combined resistance is only slightly less than that of one rod, but if they are far enough apart, the combined resistance will approach one-half of that of one rod.

Table I. Flashover of Suspension Insulators With 10-Inch Diameter and $5^3/_4$-Inch Spacing

Barometer 30 Inches, Temperature 77 Degrees Fahrenheit, Vapor Pressure 0.6085 Inch

Number of Insulator Units	60 Cycles RMS Dry Kv	60 Cycles RMS Wet Kv*	Impulse Kv
1	80	50	125*–150
2	155	90	250*–255
3	215	130	355
4	270	170	440
5	325	215	525
6	380	255	610
7	435	295	695
8	485	335	780
9	540	375	860
10	590	415	945
11	640	455	1,025
12	690	490	1,105
13	735	525	1,185
14	785	565	1,265
15	830	600	1,345
16	875	630	1,425
17	920	660	1,505
18	965	690	1,585
19	1,010*	720	1,665
20	1,055*	750	1,745
21	1,100	780	1,825
22	1,145	810	1,905
23	1,190	840	1,985
24	1,235	870	2,065
25	1,280	900	2,145
26	1,325	930	2,225
27	1,380	960	2,305
28	1,410	990	2,385
29	1,460	1,020	2,465
30	1,505	1,050	2,550
31	1,550	1,080	2,630
32	1,600	1,110	2,715

From reference 7, except as otherwise noted. Impulse values based on 1.5x40 microseconds positive wave. Values for 21 units and more are extrapolated.

* From Locke Insulator Corporation catalog.

In a single circuit line of the usual design, with the conductors horizontally arranged (Figure 5), it works out well mechanically to place the ground wires above the tower legs or above the poles in a wooden H-frame arrangement, that is, on a vertical plane about midway between the two conductors. This has proved in practice to give good protection.

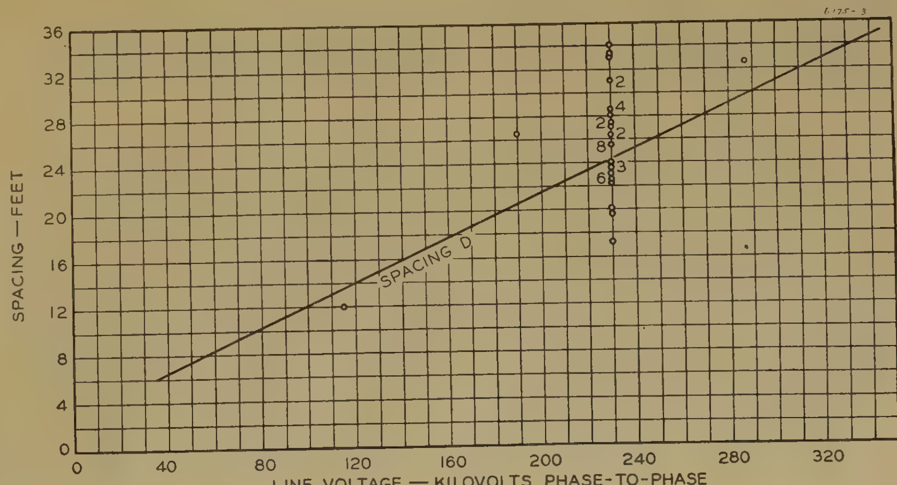
In the double circuit vertical arrangement (Figure 6), the ground wires are well coupled with the top conductors but are more remote from the middle and especially the bottom conductors, so that the coupling factor is rather low. To offset this loss of coupling factor, the ground wires should be spread farther apart. A good position in practice has been found to be about directly above the outermost conductors, which are usually the middle conductors.

The ground wires should be high enough above the conductors to avoid flashover between conductors and ground wires at midspan. However, if the ground wires are too high, the coupling with the conductors will be small and in case the tower voltage is raised above ground by lightning current in the tower, the voltage across the insulator string will be greater than it would have been with the ground wires closer to the conductors.

It has been found by experience that a height of approximately two-thirds the spacing between conductors for the horizontal arrangement and approximately equal to the spacing between conductors for the vertical arrangement is about right to give good coupling to the conductors, and at the same time avoid midspan flashover. Spacing C, Table IV, follows these criteria.

In regions where snow, ice, or sleet are likely to accumulate on the conductors, it happens occasionally on vertical arrangements (Figure 6) that the bottom conductor unloads its snow and ice and swings up level with or above the middle conductor. Also, in case of a coating of ice or sleet and a light wind in the proper direction, sometimes dancing or galloping occurs. In this case, the conductor swings in an oval loop (Lissajous figure). Lightning is not apt to be present under these conditions. Nevertheless, the spacing between conductors when in their closest position should be amply safe, so

Figure 3. Horizontal arrangement of conductors; actual spacing between conductors and, for comparison, spacing D for average line insulation



that the dynamic voltage or overvoltages will not flash over. It should be noted in this case that the voltage between conductors is the line-to-line voltage. Spacing E equal to two-thirds A would have a dry 60-cycle flashover approximately 2.5 times the line-to-line voltage, while spacing E equal to A would have a 60-cycle dry flashover approximately 3.5 times the line-to-line voltage. In practice, spacing E may be as small as 0.5A for favorable snow and ice regions and up to 1.5A for regions where unloading of ice or snow or dancing of conductors may be expected.

Table IV gives the various spacings which co-ordinate with the different numbers of insulator disks, all based on $5^3/_4$-inch spacing between disks. In spacing D the width of the tower leg has been neglected. These values are plotted on Figure 7. If, for example, for a given transmission line seven disks are selected, then the spacings in Table IV corresponding to seven disks will co-ordinate with this number of disks. If for the same line it is desired to use nine disks, then all of the spacings should be increased to those corresponding to nine disks.

On Figure 3 are shown spacings D based on average line insulation from Figure 2, for comparison with the spacings found in practice for horizontal arrangement of conductors. On Figure 4 are shown spacings C based on average line insulation from Figure 2, for comparison with the spacings found in practice for vertical arrangement of conductors.

On lines with pin type insulators, there will be no swing of the conductors at the structure, so that spacing A could be used between conductor and down-lead as far as electric clearances are concerned. However, mechanical considerations may call for larger spacings, and spacing B would be applicable.

Table II. Flashover of Standard Rod Gaps

Barometer 30 Inches, Temperature 77 Degrees Fahrenheit, Vapor Pressure 0.6085 Inch

Gap Spacing, Inches	60 Cycles RMS		
	Dry Kv	Wet Kv*	Impulse Kv
1	22.5	—	38
2	40	—	60
3	50	—	75
4	58	—	91–95
5	66	—	106–114
6	74	—	128–141
7	—	—	141–155
8	89	—	159–166
9	—	—	175–178
10	105	80	190
15	150	125	275
20	200	170	350
30	295	255	505
40	385	345	650
50	480	430	800
60	575	515	945
70	665	600	1,095
80	755	675	1,240
90	835	755	1,385
100	910*	835	1,530
120	1,055	970	1,820
140	1,200	1,110	2,110
160	1,330	1,230	2,400
180	1,450	1,350	2,690
200	1,560	1,460	2,980

From reference 7, except as otherwise noted. Impulse values based on 1.5x40 microseconds positive wave. Values for spacing above 100 inches are extrapolated.

* From curve of Pittsfield General Engineering Laboratory.

Wood Pole Lines

Wood pole lines with hardware bonded and connected to grounded down-leads would be treated exactly the same as the steel tower lines previously discussed. If the hardware is not bonded, there will be a certain amount of wood insulation in circuit, and the impulse value of this wood when dry may be evaluated at 100 kv per foot, and when wet at 50 kv per foot.[8,9] For conservative design, 50 kv per foot or the equivalent of one-half disk insulator for each foot of wood may be used.

On account of the possibility of wood poles and crossarms splitting from the lightning flashover, it is desirable to have the flashover go from the conductor to the down-lead through air, rather than via the insulator string and crossarm or pole. If the impulse flashover through air is made about the same as the insulator flashover, the added insulation of the wood in series with the insulator will ensure that the flashover will take place through air to the down-lead.

Tower Footing Resistance

The selection of a particular insulation and co-ordinated spacing of conductors does not assure that the line will not flash over because of lightning. Flashover can only be prevented by

1. Shielding the conductors from direct strokes by properly placed overhead ground wires.

2. Low tower footing resistance so that the product of tower current and resistance will not exceed the flashover of the insulation and spacing selected.

Lightning currents to be expected may be determined from reference 10. Experience has shown that flashover usually will not take place if the product of tower current and footing resistance does not exceed the 1.5x40 positive flashover of the insulation or spacing to ground.[11] For example, on a line with seven insulator units having an impulse flashover of 695 kv, and assuming 100,000 amperes tower current, the footing resistance would have to be reduced to about seven ohms in order to prevent flashover. Reference 10 indicates a reduction in footing resistance during surges as compared with the tested resistance. This reduction is greater the larger the lightning current and the larger the normal resistance. The reduction may be considered as a factor of safety when figuring on possible

Figure 4. Vertical arrangement of conductors; actual spacing between conductors and, for comparison, spacing C based on average line insulation

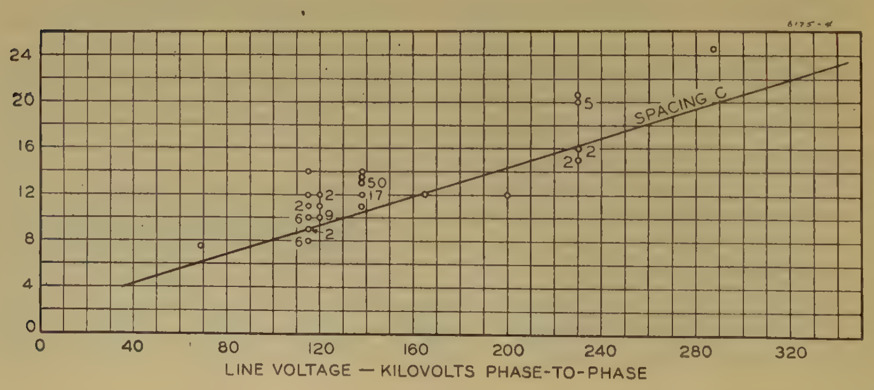
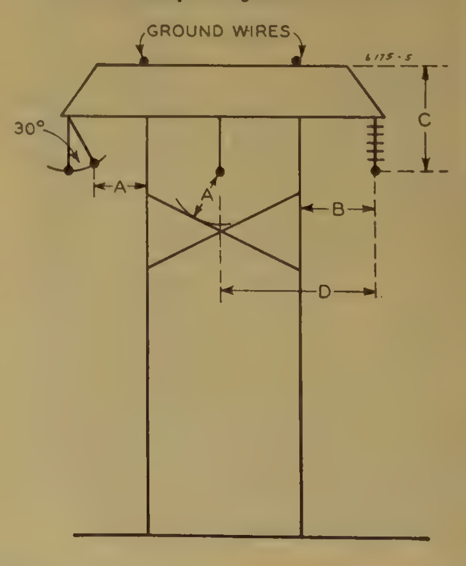


Figure 5. Single-circuit transmission line with horizontally arranged conductors



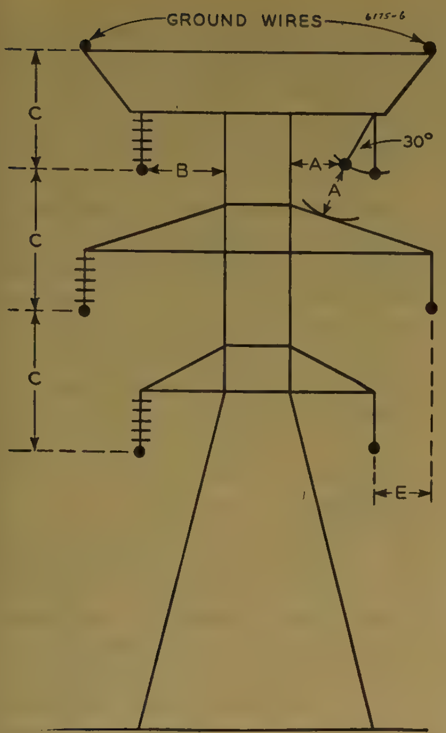


Figure 6. Double-circuit transmission line with vertically arranged conductors

Table III. Minimum Number of Insulator Disks for Transmission Lines Based on $5^3/_4$-Inch Spacing

(1) Circuit Voltage Kv	(2) Number of Disks to Have 60-Cycle Wet Flashover Equal to or Greater Than 3.5×L-N Voltage	(3) 4×L-N Voltage	(4) 4.5×L-N Voltage	(5) 5×L-N Voltage	(6) Average Practice Number of Disks	(7) Basic Impulse Level Kv
34.5	2	2	2	3	3	200
46	3	3	3	4	4	250
69	4	4	5	5	5	350
92	5	5	6	7	7	450
115	6	7	8	8	8	550
138	7	8	9	10	9	650
161	8	9	11	12	11	750
196	10	11	13	14	13	900
230	12	14	15	18	15	1,050
287	15	18	20	23	19	1,300
345	19	22	25	29	22	1,550

flashover based on the normal resistance. Also, by neglecting the coupling factor an additional factor of safety is secured.

In some locations where the ground resistance is inherently high, it may be necessary to reduce the footing resistance by driven rods or counterpoise wires. Nevertheless, it may not be feasible or economical to obtain footing resistance values low enough. In these cases, it may be desirable to supplement the means already used with ground-fault neutralizers, expulsion protector tubes, or automatic high speed reclosing. These supplementary means have been described amply elsewhere.[12,13,14]

Conclusions

1. The 60-cycle wet flashover of line insulation should equal or exceed the voltages from faults and switching surges, that is, under the most advantageous conditions from 3 to 3.5 times normal line-to-neutral voltage, and under less favorable conditions up to 5 or 6 times normal line-to-neutral voltage. The impulse flashover should equal or exceed the basic impulse insulation level for apparatus.

2. The number of insulators having a wet 60-cycle flashover of 3.5 times line-to-neutral voltage is the minimum number which will meet the basic impulse insulation level.

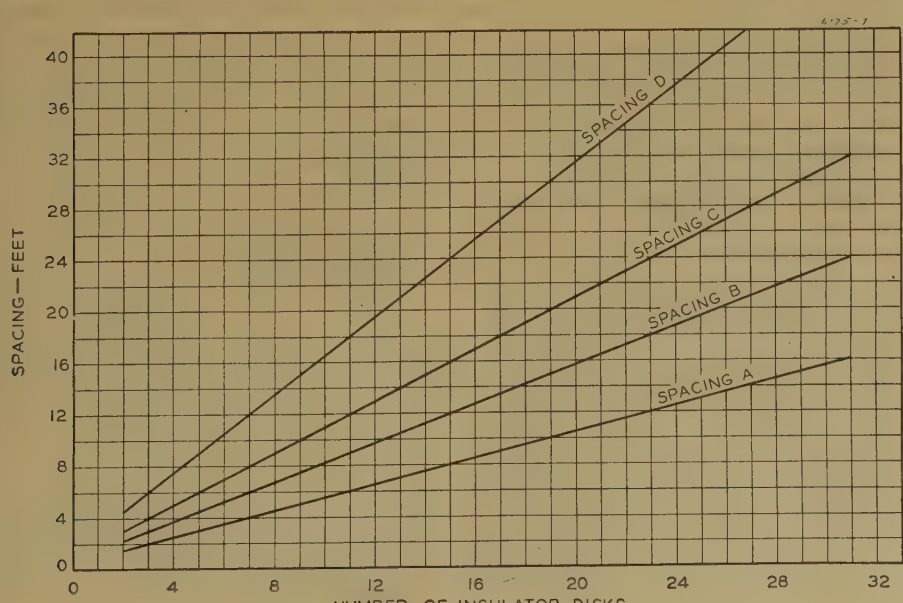


Figure 7. Relation between number of insulator disks and spacing for transmission lines:

A—Conductor to nearest tower steel
B—Conductor to vertical tower members
C—Between vertical conductors and between conductors and ground wires
D—Between horizontal conductors

3. Spacings between conductors, from conductors to tower, and from conductors to overhead ground wires should co-ordinate with the impulse flashover of the number of insulators chosen.

Spacing *A* from conductor to the nearest tower steel should have an impulse flashover about ten per cent greater than the insulator string flashover. In feet this is roughly equal to $\dfrac{\text{number of disks}+1}{2}$.

Spacing *B*, the horizontal distance from conductor to tower when the conductor is hanging vertically, equals $1.5A$.

Spacing *C*, between horizontal planes through conductors in the vertical arrangement, and between the plane of upper conductors and plane of ground wires, equals $2A$. The same spacing is used between the plane of conductors and plane of ground wires in the horizontal arrangement.

Spacing *D*, between conductors in the horizontal arrangement, is $3A$.

Spacing *E*, between vertical planes through middle and bottom conductors in vertical arrangement, varies from 0.5 to $1.5A$, depending on severity of snow and ice, and wind conditions.

4. Immunity from flashover depends not only on the insulation and spacings selected,

Table IV. Co-ordination of Spacings With Number of Insulator Disks

Number of Insulator Disks*	Spacing A, Feet	Spacing B, Feet	Spacing C, Feet	Spacing D, Feet
3	2	3	4	6
5	3	4.5	6	9
7	4	6	8	12
9	5	7.5	10	15
11	6	9	12	18
13	7	10.5	14	21
15	8	12	16	24
17	9	13.5	18	27
19	10	15	20	30
21	11	16.5	22	33
23	12	18	24	36
25	13	19.5	26	39
27	14	21	28	42
29	15	22.5	30	45
31	16	24	32	48

* Spaced $5^3/_4$ inches.

Spacings for intermediate numbers of disks may be interpolated.

Refer to Figures 5 and 6 for location of spacings *A*, *B*, *C*, and *D*.

Transmission Rating of Telephone Systems

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Synopsis: The problem of rating the transmission performance of telephone connections is considered, including a review in considerable detail of a method proposed by J. R. Hughes.[1] With this background, a modification of Hughes' technique is presented with indications of certain attendant advantages.

A TELEPHONE, by definition, is a device for projecting the human voice beyond its normal range. There are other systems, of course, such as broadcasting, public address, paging, and recording-reproducing systems which perform this same general function. Each has its own type of service, the nature of which accounts for the differences in equipment and technique peculiar to each system. The service offered by the telephone system makes it the most complex of all. Each subscriber must have his own telephone set which must have access to all other subscribers individually for 2-way conversation and be able to establish and disengage all possible interconnections. The inherent complexity of establishing the transmission paths may tend to distract attention from the primary function of the system, which is to convey intelligence by means of the spoken word. Since this is the reason for the existence of a telephone system, there should be adequate means of measuring its success in accomplishing this end. This is important both to maintain adequate service and to point the way for progress in the art.

This paper will discuss the methods of rating which have been used from time to time throughout the course of development of telephone systems. Special attention will be given to a technique recently proposed by J. R. Hughes which will be reviewed in considerable detail. With this background, a presentation will be given of a modification of Hughes' technique with certain indicated advantages.

Since this discussion is concerned entirely with the transmission problem, the terms telephone system or connection are to be understood to refer to the equipment which makes up a complete voice transmission path from one subscriber to another.

Basis of Rating Methods

Inasmuch as the fundamental function of the telephone system involves the transmission of voice-produced electric currents over a distance, the first essential for the system is satisfactory transmitting and receiving equipment. With such equipment available it is immediately apparent that the loss in the telephone line (or other connecting medium) is of primary importance since with increased loss the weaker signals, and eventually all signals received, will fall below the audible limit. For that reason, various methods for rating of telephone systems, which have been used throughout the years, use as their unit some measure of the loss in a unit length of line.

Rating Electric Circuit Components

The earliest method used for specifying line losses was to indicate the number of miles of standard cable in the connection. Where lines other than standard (19 gauge) cable were used, the number of miles of standard cable which would give the same reduction of current from input to output at an average frequency of say 800 cycles per second would be the measure of the line loss. When dealing with such lines it was necessary to use some measure of current reduction in the line to determine the length of standard cable which gave equal loss. Accordingly, a further step in development was to state the line losses directly as a function of this current reduction. Thus, the 800-cycle mile was born. It is a unit based on the loss in passing through one mile of standard cable at 800 cycles per second, which is defined mathematically as the natural logarithm of the ratio of the input to output currents in one mile of standard cable (which is 0.109). Thus the formula for determining the number of 800-cycle miles corresponding to any values of input and output current on a properly terminated line is $2.30/0.109\,[\log_{10}(I\text{ in}/I\text{ out})]$. Other circuit components are then rated in terms of the number of 800-cycle miles corresponding to the current or power reduction resulting from their insertion in the circuit.

Ratings of Transmitters and Receivers

The most difficult problem in telephone transmission rating has been the development of an adequate method for determining and expressing in simple and in-

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but on adequate shielding by overhead ground wires and sufficiently low tower footings resistance. In general, flashover will not be experienced if the footing resistance is less than the 1.5x40 microseconds positive impulse flashover of the insulation, divided by 100,000.

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formative terms the performance of transmitters and receivers.

Historically, the earliest attack on the problem is found in the volume or loudness tests. They were based on the fundamental assumption that two over-all connections which would produce equal loudness in the listener's ear were equally satisfactory and, therefore, would be given equal ratings. Observers were asked to judge the relative loudness of two such connections. A series of observations were made with successively different equivalent lengths of a standard trunk in one connection as compared to the other. When settings of trunk were determined which gave equal loudness for the two connections, the difference in equivalent length or attenuation was taken as the difference in rating of the two systems. Thus the performance of over-all connections, including the effects of various subscriber sets, was rated in terms of equivalent trunk losses.

Initially, the standard trunk used for such comparisons was the 19-gauge standard cable (referred to previously) and the ratings were expressed in terms of a number of miles of standard cable. It had the disadvantage of not producing uniform attenuation throughout the frequency range. Consequently, adjustments of the trunk changed not only the level but also the response-frequency characteristic by producing larger changes of attenuation at high frequencies than at low frequencies, thus making equality of loudness difficult to judge. It eventually was decided to replace the standard cable by a 600-ohm distortionless trunk. With the adoption of this trunk, the unit of line loss was replaced by a new unit of approximately the same magnitude, but defined as $10 \log_{10} (P\text{ in}/P\text{ out})$ and called at first the transmission unit (TU) and later the decibel (db).

Quality Differences

In the course of the loudness comparisons which were used in telephone instrument ratings, it often would be apparent to the observers that when the trunk had been adjusted for a loudness match, there was still a difference in the sound of the reproduced voice over the two systems, even when the distortionless trunk was used. Such differences commonly are described as quality differences. It seems evident that when the quality differences are large they so significantly may affect the intelligibility of speech transmitted over the system that volume rating alone becomes inadequate as a performance index.

Judgment Tests

Earliest attempts to take account of quality differences were the judgment tests. In these tests after the systems under comparison had been adjusted for equal volume, various observers were asked to listen to representative sentences transmitted over the system and give their judgment as to which system they preferred. The systems were rated according to the relative number of votes cast for each. A glaring weakness of the method is that the preferences expressed often may be quite unrelated to the actual transmission merits of the system and thus may be definitely misleading.

Articulation Testing

A much more fundamentally sound and informative method of evaluating the effects of quality differences is the articulation test. It has been used since before 1910[2] and its value still is recognized. The test is conducted in the following manner. Each of a number of callers speaks a list of meaningless monosyllables or "logatoms" over each of two systems to be compared. The logatoms preferably are inserted in sentences. A group of observers at the receiving end of the systems record the logatoms as they understand them. The percentage of sounds or logatoms correctly received is called the per cent articulation of the system. Per cent articulation is more significant than volume rating alone in that it does take account of quality differences as well as volume differences. A definite disadvantage, as compared to volume ratings, is that ratings of system components may not be simply combined to determine over-all ratings of connections of varied make-up. It is conceivable, however, that series of articulation tests may be taken on each system in such a manner that their results might finally be expressed in units suitable for this purpose. This will be discussed further in connection with immediate appreciation testing.

Analysis of Quality Differences

The existence of quality differences between systems led to development of other kinds of testing in which the specific objective was to explain the quality differences. Such tests are referred to commonly as objective tests, since they measure the physical characteristics of the system without dependence on the user's judgment or opinion. Their highest development is found in the use of artificial mouth and ear equipment.[3] They are very valuable tests for their intended purpose of analysis of performance. However, they do not afford a straightforward means for the rating of over-all performance in simple terms which are interpreted readily and hence need not receive further consideration here.

The Complete Rating Problem

The preceding discussions have called attention to the fact that one telephone connection may differ from another in the quality of the received speech as well as in volume or loudness. Quality differences may be either in frequency distortion (unsatisfactory balance of response throughout the voice frequency range) or in nonlinear distortion (the introduction by the circuit equipment of additional frequencies not present in the original speech). In addition to these effects, there are differences caused by the fact that the circuit is required to provide 2-way transmission with an invariable circuit, that is, without switching of transmitter and receiver alternately into the circuit for talking and listening, respectively. The result is the condition commonly described as side-tone, that is, a reproduction of a portion of the sound picked up by the transmitter of a given set in the receiver of the same set. During transmitting the presence of this side-tone causes the user to lower his voice, the amount depending on the amount of such side-tone. On receiving, side-tone introduces noise into the listener's receiver picked up from the room in which he is listening, the amount of such noise being dependent on the amount of room noise and the efficiency of the side-tone path.

It is evident that any figure of merit which is intended to serve as a measure of the over-all value of a telephone connection in conveying intelligence in actual service must be obtained in such a way as to include the effects of all of these characteristics of the equipment and their interactions on the users.

The Repetition Rate Test

The repetition rate test[4,5] was conceived and developed in the Bell Telephone System as a practical means of providing such a figure of merit. In this method, the test data are obtained by monitors on actual telephone connections of known make-up who make records of the number of times that the users, in the course of their conversations, give evidence of failure to understand each other. The monitors also observe the elapsed

time for each conversation. The data are summarized to give the number of repetitions required per unit time. This is called the repetition rate for the connection. Such data are statistical in nature and a large number of such observations must be made to give results of acceptable precision. Data are taken on a large number of different connections to cover the range of equipment types to be rated. The repetition rate is taken as the fundamental measure of the degree of success in carrying on conversations. Advantages of the volume rating system are retained by including in these tests measures of the variation of repetition rate with variation of length, or attenuation in decibels, of distortionless trunk included as part of the connection for this group of tests. It is then possible to convert variations of repetition rate caused by changes of any part or component of a telephone connection to equivalent changes in distortionless attenuation expressed in decibels. This gives a rating system in which the performance of any component of a connection may be stated in terms of the number of decibels of effective transmission loss as compared with the corresponding component of a standard reference system.

The basis of these ratings, the repetition rate test, has come to be generally regarded as the ultimate criterion of satisfactoriness of the over-all transmission path of a telephone connection. Disadvantages of this method are that it is essentially a field test, not adaptable to the laboratory, requires a very long time to obtain data, and is not generally practical for operating companies or manufacturers to use because of privacy requirements. The Bell System was able to overcome the latter difficulty by using several hundred stations in the American Telephone and Telegraph Company headquarters building and a like number at the Bell Telephone Laboratories.

The Immediate Appreciation Test

The immediate appreciation test[1] is an attempt to bring the repetition rate test into the laboratory. In other words, it is a proposed means of overcoming the disadvantages of the repetition rate test and still obtaining a rating which will include the effect of all the various system characteristics and the user's reactions to them.

In the application of the method, talking and listening crews speaking over the connection to be rated use lists of 20 sentences, each of about six to ten words, simple and straightforward in type, and extracted from the ordinary news columns

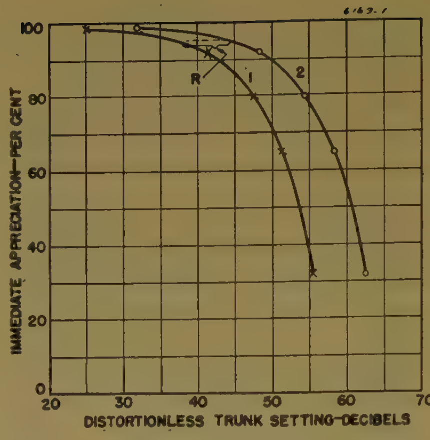

Figure 1. Immediate appreciation test—comparison of transmitting substations

Curve 1—Connection from transmitting substation *A* to receiving substation

Curve 2—Connection from transmitting substation *B* to receiving substation

of a daily newspaper. The listener is simply asked to decide whether or not he *appreciated* the meaning of each sentence *immediately* without any sort of delay or introspection on his part. On this basis he records each sentence as a success or failure. The percentage of successes is the percentage immediate appreciation and is taken as the criterion of performance of the system. It is this particular testing technique which distinguishes the test from other subjective tests. (These are tests which use talking and listening crews.) Other details which will be discussed have to do with means of taking account of the effect of service conditions (that is, speaker's volume level and room noise) and the means of presenting the data and converting it to more useful form.

Assuming the tests are so arranged as to take adequate account of service conditions, it will be seen that the outstanding difference from the repetition rate test is in the fact that the listeners also become observers in the test, so that the test data come directly from them instead of from monitors. The record, however, is based on fundamentally the same reaction, the difference being that the listener, as a result of a failure to appreciate the meaning of what has been transmitted, records this failure instead of requesting a repetition. Advantages are that the tests can be made in the laboratory, the required data can be obtained in a much shorter time since continuity of testing may be obtained, and the sensitivity of the method is higher, thus reducing the amount of required data and testing time. A possible disadvantage is a certain degree of artificiality, a fault of any laboratory test. In this respect, however, the immediate appreciation method has considerable advantage over the articulation test in which meaningless monosyllables constitute the matter transmitted. Here again there is a considerable advantage in time for the appreciation test since consistent results are found to be obtainable with much less matter transmitted. There is also the advantage of practically unlimited source of material in contrast to the limited number of logatoms available, thus reducing practice effects caused by familiarity with the material transmitted. The significant point, however, in the appreciation tests is the fact that the records obtained are intended to indicate specifically and directly the success of the telephone connection in conveying the meaning of connected words (that is, expression of thought) from talker to listener. Aside from these disadvantages of the articulation method, there appears to be no reason why the methods to be presented for rating of systems by application of the immediate appreciation technique might not use equally well the logatom articulation technique.

In the immediate appreciation tests, the direct result of a test run on two systems to be compared will show for the particular test conditions the per cent appreciation for each system. This, in itself, is a valuable indication of the relative performance of two systems. However, in order to enable expression of the results in more useful form, a number of runs are taken on each system, the amount of distortionless trunk in the connection being set at a new value for each run. Resultant data permit the plotting for each system of a curve of per cent appreciation as a function of trunk attenuation. Such curves are shown in Figure 1. Figure 2 shows by block diagram typical circuits which are used in obtaining these data. In Figure 1 the difference R of attenuation in decibels at a given level of per cent appreciation is taken as the difference in effective transmission of the two systems for the given test conditions. Thus, if this per cent appreciation is taken at a value which is agreed upon as representing a satisfactory standard of communication efficiency, it is indicated in the given example that 7.5 decibels more attenuation may be inserted in connection 2 than in connection 1, and still furnish transmission equal to the specified standard. It might be argued that 100-per-cent appreciation should be used. However, for practical reasons, a value somewhat short of this (95 per cent) is used. One reason is that the slope of the curves in this region allows the comparison to be made

more precisely; also, in general, it is not economical to design for perfection.

This method of interpretation of results assumes that it always will be possible to obtain a curve of the type shown in the figure for any system. A little consideration will show that it is a fundamental result of the test method. A difficulty, of course, would be presented by a system of such poor characteristics that 95-per-cent appreciation would be impossible to attain at any attenuation setting. However, there would be no practical interest in trying to rate such a system.

Treatment of Service Conditions

In his article on this method, J. R. Hughes gives extensive discussion of a statistical technique for taking account of the range of service conditions. The statistical method is based on frequencies of occurrence of various values of speaker's volume level and room noise determined from a large number of measurements of these quantities (in the case of room noise at actual subscriber locations). It consists of a procedure for weighting the observed data in accordance with these frequencies of occurrences. In making these tests the talker's volume level has to be treated in such a way that any tendency of the presence of side-tone, which he hears from his own receiver, to influence the loudness of his voice will be permitted to have its normal effect. This means that no artificial control should be exercised, such as the instruction to maintain a constant specified speaking level which is the common procedure in the usual voice-ear volume tests or articulation tests. In lieu of such control, the method of relating the speaking levels of the crew to the range of service distribution is by measurement of their natural speaking level; that is, the level at which they speak most naturally when conversing, without a telephone, in a quiet room.

In obtaining the data for application of this statistical technique, it is necessary to take several families of curves to give the required information on variation of per cent appreciation with service conditions. The several families are taken each for a different fixed value of room noise at the transmitting end of the connection, each curve in a given family being obtained for a different value of room noise at the listening end. Thus each individual curve of each family gives the variation of per cent appreciation with attenuation of distortionless trunk for a different combination of transmitting and receiving end room noises. About four families of four curves each are required. Variations of speaking level are taken as equivalent to variations of distortionless trunk setting.

By application of the statistical technique to these data, a single figure of merit is obtained which indicates how much more attenuation may be inserted in the better of two circuits and still obtain a grade of transmission at least equal to the accepted limiting value (95-per-cent appreciation) on the same percentage of all calls as on the poorer system.

A Modified Technique

In the foregoing discussions of immediate appreciation testing, interpretations of the test data have been made in terms of the excess amount of distortionless attenuation which must be inserted in the better of two systems in order to make the two systems equivalent in their ability to transmit intelligence. In the data discussed in connection with the statistical technique, this equivalence is established on the basis of an equal per cent of calls, enjoying at least 95-per-cent immediate appreciation; and the resultant rating is a composite figure dependent on all combinations (suitably weighted) of room noise levels and speaking levels occurring in service. In contrast to this, in standard Bell System effective ratings, the equivalence of two systems is established on the basis of equal success in conveying intelligence in the presence of a single representative (reference) value of room noise existing at each end of the connection in each of the two systems. It seemed desirable in developing a laboratory test for rating telephone systems in this country to express the ratings on this basis.

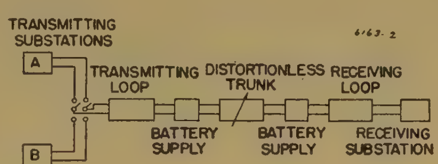

Figure 2. Diagram of connections for a typical immediate appreciation test

The adoption of such a basis would result, also, in a very considerable reduction (to about 1/16) in the amount of test data required. This is true since it would mean that the data for determination of the several families of curves discussed in connection with the statistical technique would be reduced to that necessary to give a single curve for each system under test, such as 1 or 2 shown in Figure 1. The full extent of this simplification is not at once evident and may not be permissible always, since the technique employed still must take account of the service range of speaker's volume level, including the effects of side-tone characteristics of the transmitting subsets on this level. Accordingly, these tests must be conducted without instruction to the talkers to maintain any specified speaking level.

In general, it would be necessary to obtain a family of curves of per cent immediate appreciation as a function of distortionless trunk setting for each connection to be rated where the several curves of a family would be taken, each with a different crew having a different natural speaking level. The difference R (Figure 1) between the curves for two systems being compared would be taken for each crew. A weighted average value of R then would be obtained in which the weighting would be in accordance with the frequency of occurrence of the respective values of natural speaking level in service.

The ultimate simplification possible in the use of this method results by applying the fundamental assumption which provides the basis for Hughes' statistical technique. This is derived from his observation that under actual talking conditions amplitude distortion, even in systems using carbon transmitters, is negligible. The corollary to this is that changes in the amount of attenuation in the distortionless trunk may be taken as representing, also, changes of the speaker's volume level. Applying this assumption, for example, to the curves of Figure 1 leads to the following conclusions. Suppose that curves 1 and 2 and the resultant difference R between the two systems have been obtained with a crew having a given natural speaking level. Now suppose that this data were repeated with a crew having, for example, a 10-decibel higher natural speaking level. Application of the foregoing principle leads immediately to the prediction that the new curve 1′ (not shown) for the connection involving transmitting substation A will be obtained from curve 1 by adding ten decibels to the abscissa for each point, that is, the new curve will be displaced horizontally ten decibels to the right from curve 1. The new curve 2′ (not shown) for the connection using transmitting substation B will be obtained in an exactly similar manner from curve 2. Thus the difference R' between curves 1′ and 2′ will be identical to the difference R. Hence the difference R will be independent of the natural speaking level so that

only a single curve of per cent appreciation as a function of distortionless trunk setting need be taken for each connection to be rated.

When the assumption of equivalence of variations of attenuation of distortionless trunk and of speaker's volume level are not substantiated, the method of using several crews having different natural speaking levels should be adopted. This still would require only about one-fourth of the data required by Hughes' complete statistical technique and has the further advantage in this case of taking account of any existing amplitude distortion effects which are neglectedby the fundamental assumption of the Hughes' method.

If ratings are desired for other room noise levels, they are obtained by repeating tests of the foregoing type with the room noise adjusted to the required level. As a simplifying approximation, such tests may be taken on a representative system, and the difference between the ratings for average room noise and the required room noise may be applied as a correction to the rating determined on any other system in question at the average (reference) room noise level.

Conclusion

The simpler form of the modified testing technique described has had some application in the rating of transmission performance of subscriber's sets. The limited amount of testing which has been done up to this time indicates this technique to be quite feasible. The immediate appreciation test in this general form, therefore, gives promise of being a very valuable tool in evaluating the overall merit of developments in telephone transmission equipment.

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Carrier Supervisory Control of Pumping Station Over Power Cable

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THE PROCESSING of starch requires large quantities of water. Thus, when the United States Sugar Corporation recently added the processing of starch to their activities, it was natural to install their new plant, which makes use of sweet potatoes as the raw product, at Clewiston, Fla. Clewiston is located in the Everglades near Lake Okeechobee. Lake Okeechobee is a shallow lake but it covers an area of over 700 square miles, second in area only to Lake Michigan in the United States.

When the new starch plant was planned, it was calculated that the maximum daily demand for water would be 4,000,000 gallons with 3,000,000 of these gallons requiring purification before being used in starch processing. An analysis of the water of Lake Okeechobee indicated that the best source of water for the processing of starch on a purity and temperature basis was four miles from the levee of the lake. These findings resulted in the decision to build a concrete pumping station in Lake Okeechobee, four miles from the levee and a total of seven miles from the water treatment plant where the water is treated before being used in the starch processing.

A view of the pumping station is shown in Figure 1. The inside diameter of the station is 40 feet and the walls are $2^1/_2$ feet thick. The station is slightly over 40 feet from top to bottom with the water level normally being about 15 feet above the lower floor level.

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A 24-inch-diameter cast iron pipe carries the water from the pumping station to the water treatment plant. The pipe is buried approximately four feet below the bed of the lake. A 75-horsepower electric motor and a 180-horsepower gasoline engine are used to pump water through the pipe. Electric power is fed to the pumping station from the water treatment plant over a 4,000-volt power cable which is laid in the same trench with the pipe.

The relative inaccessibility of the pumping station led to the decision to provide control of the pumping equipment from the water treatment plant rather than having an operator located in the pumping station.

The remote control of the pumping station could have been accomplished by means of wires in a submarine control cable which would have had to be installed in addition to the power cable, or by means of power-line carrier supervisory control operated over the power cable. Economic considerations led to the application of carrier equipment to the 4,000-volt power cable.

Figure 2 is a single line diagram of the essentials of the power system at the water treatment plant and at the pumping station. The connections of the carrier equipment to the power cable also are indicated.

No carrier line traps were necessary on this installation because the power cable is terminated with the electric motor at the pumping station and with an autotransformer at the water treatment plant.

New Features of Installation

This installation of carrier-operated supervisory control equipment incorporates quite a number of operating and design features not previously employed. Following is a list of these features:

1. Operation of power-line carrier equipment over a submarine cable.
2. The use of standard low voltage capacitors for coupling to the power line instead of standard coupling capacitors, thus making it possible to omit line tuning units.
3. The complete incorporation of power-line carrier equipment including coupling equipment in a metalclad switchgear assembly.
4. Selective telemetering through the

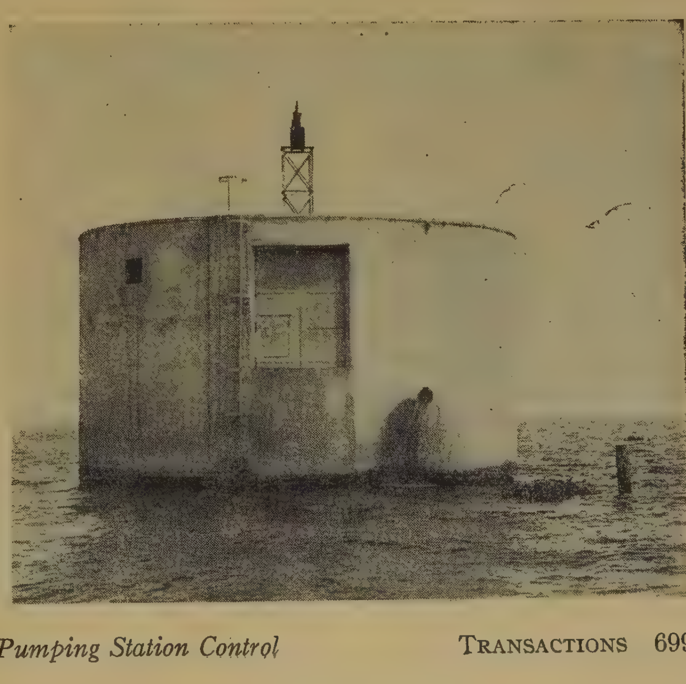


Figure 1. A view of pumping station showing entrance door

supervisory control equipment over a single-frequency power-line carrier channel without the use of audio tones for modulation of carrier.[1]

5. Establishment of communication circuit over single-frequency power-line carrier channel through supervisory control equipment.

Cable Characteristics

A total of 35,000 feet of 3-conductor power cable is used between the water treatment plant and the pumping station for the transmission of 4,000-volt 3-phase power.

A 22,800-foot length of armored submarine cable is used from the levee on the shore to the pumping station. The armored cable is laid in the same trench with the 24-inch water pipe. The submarine cable has number 4 conductors and is protected with number 8 Birmingham wire gauge galvanized armor. The outside diameter of this cable is approximately 2.06 inches. Figure 3 shows the construction of the armored cable.

A 12,200-foot length of rubber-jacketed cable is used from the levee, where it is spliced to the submarine cable, to the water treatment plant. This cable is laid in trenches across farm land at a depth of approximately four feet. No cover protection is used as this is below cultivating depth.

The characteristic impedance of ordinary high-voltage transmission lines is about 800 ohms for line-to-line channels and about 500 ohms for line-to-ground channels.[2] The characteristic impedance of both the armored and the jacketed cable used on this installation is approximately ten per cent of the usual values for ordinary transmission lines, the line-to-line impedance being approximately 85 ohms and the line-to-ground impedance being approximately 45 ohms.

When the carrier equipment was first placed in operation, the carrier sets were used to obtain a check on the line-to-ground attenuation values for the cable, which had been calculated previously from factory test data.

To obtain data necessary to check the factory test data, the carrier equipment at each location was set on the proper tap of the matching transformer to match the 45 ohms characteristic impedance of the cable. The accuracy of calculated attenuation values is dependent on having the line terminated in its characteristic impedance. The complex nature of the carrier receiver circuit makes it impossible to choose a setting which will provide exactly the right termination for all frequencies. However, because of the relatively high cable losses it was believed that only slight inaccuracies would result from the termination not being exact at all frequencies. The results obtained indicate that the carrier equipment provided satisfactory termination at all frequencies.

The carrier transmitter at the pumping station was adjusted to various frequencies with the high frequency voltage and

Table I. Calculated Decibel Loss for 4,000-Volt Power Cable (22,800 Feet Armored and 12,200 Feet Rubber Jacketed) From Field Data*

Frequency, Kc	Current Input	Voltage Input (E_1)	Voltage Output (E_2)	Impedance, Ohms	Total Decibel Loss	Decibel Loss Per Mile
175	0.515	22.8	0.38	44.3	35.56	5.37
165	0.510	23.2	0.58	45.5	32.04	4.83
150	0.520	23.3	0.79	44.6	29.40	4.43
135	0.520	23.6	1.02	45.4	27.30	4.12
120	0.515	24.5	1.30	47.6	25.50	3.85
105	0.490	23.7	1.95	48.3	21.68	3.28
90	0.510	24.5	2.85	48.0	18.68	2.82
75	0.530	26.4	4.20	49.8	15.96	2.41
60	0.540	26.2	5.50	48.5	13.56	2.05
50	0.535	28.1	7.07	52.5	12.00	1.81
40	0.560	26.2	8.40	46.8	9.88	1.49

* Decibel loss $= 20 \log_{10} \frac{E_1}{E_2}$.

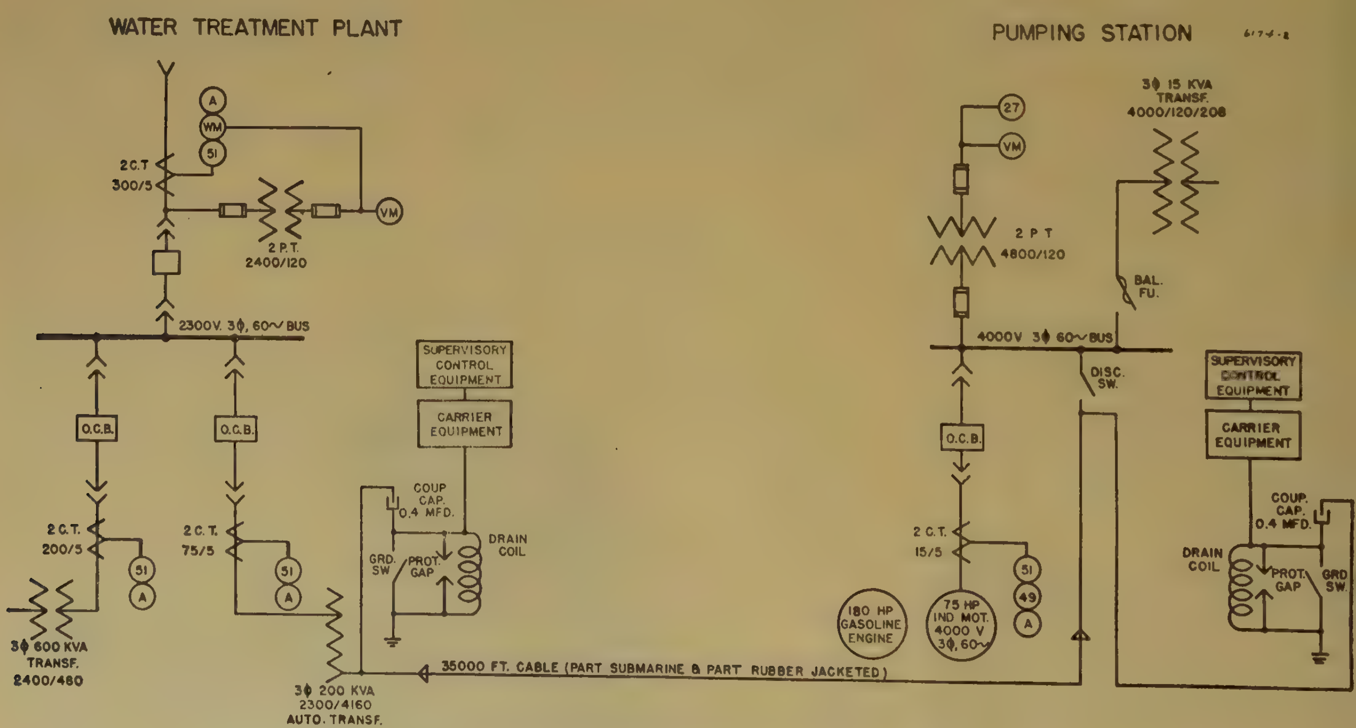


Figure 2. Single line diagram of power system at the water treatment plant and at the pumping station

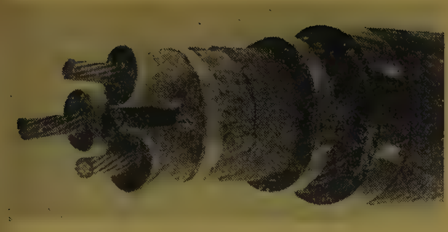


Figure 3. Cutaway view of armored submarine cable

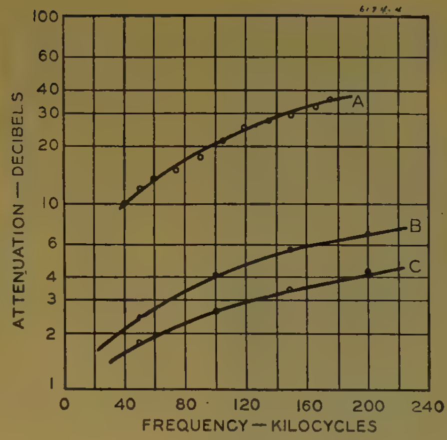


Figure 4. Curves of cable attenuation

A—Total decibel loss for entire cable from field data
B—Decibel loss per mile of submarine cable from factory test data
C—Decibel loss per mile of jacketed cable from factory test data

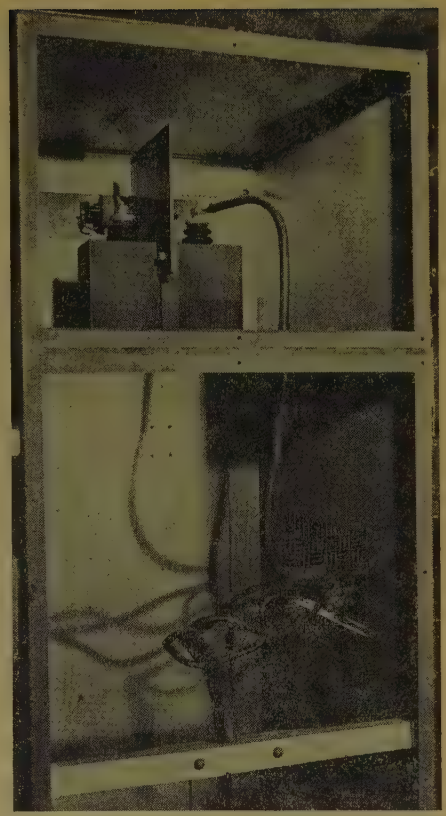


Figure 5. Rear view of metalclad unit with cover removed showing coupling capacitor

current values being read for each different frequency. The output voltage also was read at the water treatment plant for each different frequency.

From the test data obtained it was possible to calculate the attenuation values of the cable at the various frequencies. Transmission losses usually are expressed in decibels. The decibel is expressed numerically as follows:[3]

$$db = 10 \log_{10} \frac{P_1}{P_2}$$

where

db = decibel loss (attenuation)
P_1 = power in
P_2 = power out

If the two points at which the voltage is measured have identical impedances, the decibel also can be expressed as follows:[3]

$$db = 20 \log_{10} \frac{E_1}{E_2}$$

where

db = decibel loss (attenuation)
E_1 = voltage input
E_2 = voltage output

Table I shows the readings of current and voltage as well as the calculated values of decibel loss and impedance. The values of impedance which were calculated from the voltage and current input readings indicate that the termination consisting of the carrier receiver at the water treatment plant was not exactly correct for all frequencies since the calculated impedance values should have been the same at all frequencies.

Curve A of Figure 4 shows the decibel loss for the entire cable plotted against frequency, the values used being those shown in Table I. The values for the entire cable check well with factory test data obtained prior to installation of the cable. The factory tests indicated somewhat higher losses for the submarine cable alone and somewhat lower values for the jacketed cable alone. Curve B of Figure 4 shows the decibel loss per mile of the submarine cable and curve C of Figure 4 shows the decibel loss per mile of the rubber jacketed cable. Curves B and C are plotted from factory test data.

Carrier Coupling Equipment

A standard line of coupling capacitors is available for introducing carrier frequencies to power lines.[2] This line of coupling capacitors is available for system voltages from 15 kv to 345 kv. The desirability of keeping these coupling capacitors to a reasonable size resulted in their being designed with low values of capacitance. These low values of capacitance result in a relatively high capacitive reactance for even the carrier frequencies from 45 to 165 kc. Since the carrier sets are designed for matching a line impedance which is essentially resistive, a line tuning unit is used with the standard line of coupling capacitors to cancel the capacitive reactance of the coupling capacitor, thus making it possible to match the carrier sets to a resistive impedance.

The smallest carrier coupling capacitor which could have been used on this installation is for use on 15-kv lines. The size of this unit is such that it could not have been incorporated conveniently in the assembly of metalclad switchgear. Also, a line tuning unit would have been required since the capacitive reactance of this coupling capacitor at a frequency of 50 kc is 320 ohms whereas the characteristic impedance of the cable from line-to-ground is only 45 ohms.

Since the carrier equipment on this installation was to be operated from line-to-ground, it was decided to use a standard 2,300-volt capacitor similar to those used for power factor correction to couple the carrier equipment to the power cable. The capacitor chosen has a capacity of 0.4 microfarad. Due to the small size of this capacitor no difficulty was encountered in mounting it in the rear compartment of the metalclad switchgear unit which houses the supervisory control and carrier equipment. Figure 5 shows the capacitor mounted in the metal clad unit at the water treatment plant.

The capacitive reactance of the 0.4-microfarad capacitor which was used is only eight ohms at a frequency of 50 kc. This low value of capacitive reactance made it unnecessary to use a line tuning unit. The use of a 0.4-microfarad capacitor on this low voltage line results in a flow of less than 0.4 ampere of 60-cycle current to ground.

The carrier drain coil, grounding switch, and protective gap as shown in Figure 2, are all mounted on a protective panel mounted on the barrier between the high and low voltage compartments of the metalclad unit. The front of this protective panel can be seen near the top of Figure 6.

Carrier Equipment

The power-line carrier equipment on this installation is of the rack and panel

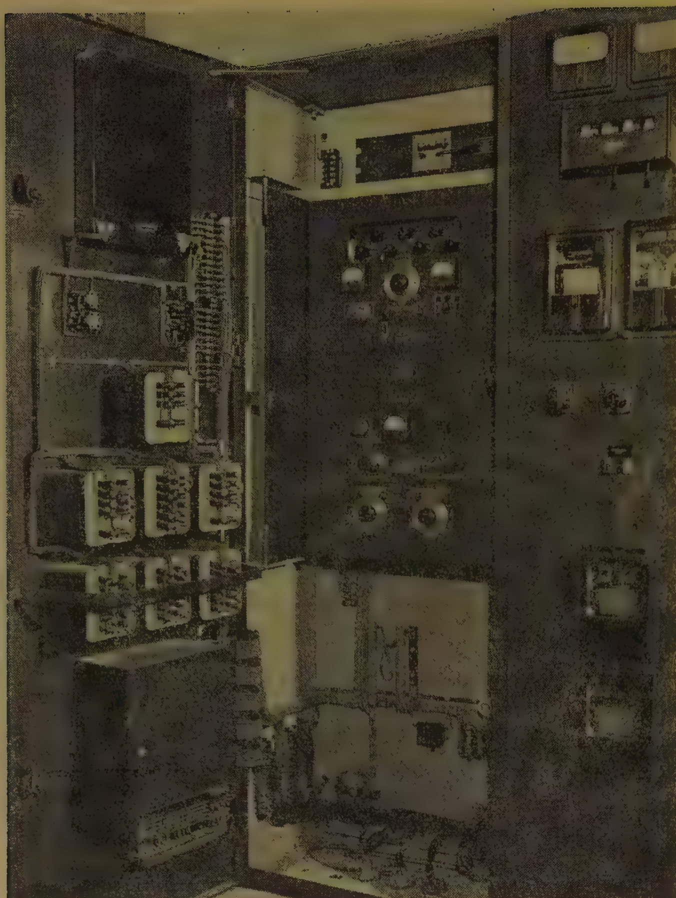


Figure 6. View at pumping station showing front of swinging carrier panel

All carrier impulses which are originated at either the pumping station or the water treatment plant result in the carrier receiver relay at both locations being operated since both transmitters and receivers operate on the same frequency. The carrier receiver relay is of the same general type as the rest of the supervisory control relays and is mounted in the supervisory control relay case.

Figure 7 shows the metalclad switchgear unit on which the supervisory control equipment is located at the water treatment plant. This figure shows the individual escutcheons or key and lamp plates for each of the functions performed by the supervisory control equipment as well as the supervisory control relay case and the water flow indicator.

The supervisory control equipment performs the following functions:

1. Start, stop, and supervise electric motor.

2. Start, stop, and supervise gasoline engine.

3. Control and supervise automatic–manual transfer.

4. Supervise electric motor annunciator.

type.[4] This type of construction makes it possible to build a unit for any carrier application, by making an assembly of the necessary standard panels. In this case a transmitter panel, receiver panel, and modulator panel were required at the pumping station and at the water treatment plant.

The carrier equipment at both locations is mounted on a swinging panel inside of one of the metalclad switchgear units. Figure 6 shows the front of this swinging panel at the pumping station.

The carrier equipment is operated from 125-volt batteries at each location. The carrier transmitters and receivers are adjustable over a frequency range from 45 to 165 kc. The transmitters have a nominal power output of 15 watts and operation is possible with the signal attenuated 30 decibels.

The transmitter panel consists of a Colpitts oscillator (one tube) and of a 6-tube push-pull amplifier. The amplified carrier frequency output is fed into an iron-cored radio frequency transformer whose secondary winding couples the carrier to the line through the coupling capacitor.

The receiver panel consists of a biased detector which is saturated when a satisfactory signal or a stronger than satisfactory signal is received. Thus, the plate current of the receiver is not materially altered for wide variations in received signal strength. The output of the detector supplies rectified current for the supervisory control receiver relay.

The modulator panel includes the matching transformers, resistors, and relays for modulating the transmitter for voice communication, for matching the carrier receiver to the telephone receiver, and for inserting attenuating resistance in the receiver circuit for improved communication.

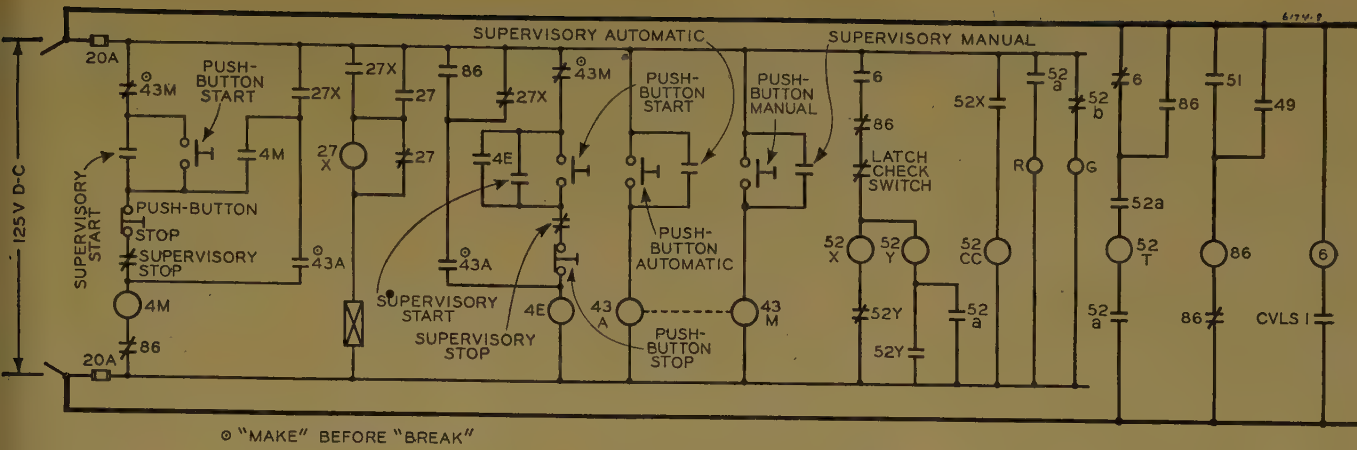
On this installation both transmitters and both receivers are operated at the same frequency. The equipment is being operated at a frequency of 45 kc, since the attenuation in the cable is considerably less at lower frequencies.

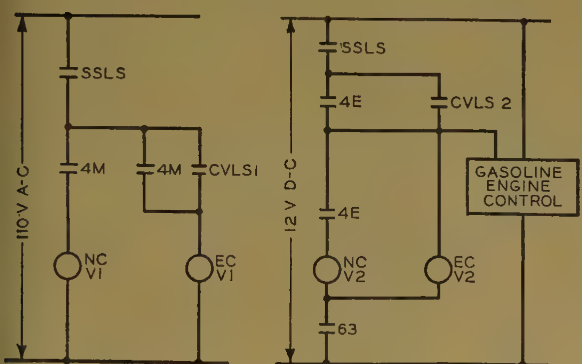
Supervisory Control

The supervisory control equipment on this installation operates by means of coded impulses with each code transmitted being checked by a duplicate code originating at the other station. The impulses are transmitted at the rate of 15 per second and all selection and control impulses are of the same duration.

Carrier impulses are transmitted by the supervisory control equipment by having the supervisory control impulsing contact close the carrier keying circuit.

Figure 7. Front view of carrier supervisory control unit at water treatment plant





Number	Device	Number	Device
4E	......Gasoline engine control relay	52T	......Trip coil for device 52
4M	......Electric motor control relay	52X	......Closing relay for device 52
6	......Electric motor starting relay	52Y	......Cutoff relay for device 52
27	......Low-motor-voltage relay	63	......Oil pressure switch
27-X	......Auxiliary low-motor-voltage relay	86	......Lockout relay
43	......Automatic–manual transfer relay	CVLS1	......Cone valve 1 limit switch
49	......Motor thermal overload relay	CVLS2	......Cone valve 2 limit switch
51	......Motor overcurrent relay	ECV1	......Emergency cone valve 1
52	......Electric motor oil circuit breaker	ECV2	......Emergency cone valve 2
52a	......Breaker auxiliary switch which is closed when breaker is closed	NCV1	......Normal cone valve 1
		NCV2	......Normal cone valve 2
52b	......Breaker auxiliary switch which is closed when breaker is open	SSLS	......Surge suppressor limit switch
		G	......Green lamp
52CC	......Closing coil for device 52	R	......Red lamp

Figure 8. Schematic diagram of automatic and manual control of electric motor and gasoline engine

5. Supervise gasoline engine annunciator.
6. Supervise burglar alarm.
7. Selective telemetering of rate of water pumped.
8. Establishment of communication circuits.

Communication

Communication between the water treatment plant and the pumping station is provided over the carrier channel through the supervisory control equipment. Picking up the telephone hand set at either location results in the proper selection codes being transmitted by the supervisory control equipment to ring the telephone bell at the location being called. The telephone bell continues to ring until the call is acknowledged by picking up the telephone hand set at the called location or until the telephone hand set is restored at the calling location. When the telephone hand set is picked up at the called location, the supervisory control equipment automatically locks out and the communication circuits are established. Push-to-talk communication is used since the carrier transmitters and receivers all are adjusted to the same frequency. When communication is established the carrier receivers are operated automatically on the linear portion of the detector tube characteristic curves so that undistorted demodulation can be effected.

The supervisory control equipment remains locked out until the telephone hand sets are restored. When both telephone hand sets are restored, the supervisory control equipment automatically unlocks and is ready immediately for other operations.

Telemetering

A recorder with a 7-day chart is installed in the pumping station to record the rate of water pumped in millions of gallons per day. This recorder operates on impulses received from a telemetering transmitter arranged to measure water flow by a venturi tube connection. The telemetering transmitter which is used to send impulses to the recorder also is connected to the supervisory control equipment so that a selective indication of rate of water pumped can be obtained at the water treatment plant over the supervisory control channel.

The telemetering equipment operates on the duration of impulse principle. Each cycle of the transmitter is divided into an on and an off period. Minimum reading is obtained on the receiver with minimum on period and maximum reading is obtained with maximum on period. The telemetering equipment used on this installation operates on a 5-second cycle. At zero water flow there is an on period of approximately 0.5 second and at a flow of water equal to full scale deflection on the receiver (15,000,000 gallons per day) there is an off period of approximately 0.5 second.

When the telemetering point is selected at the water treatment plant, the telemetering receiver is connected to the carrier receiver at the water treatment plant and the telemetering transmitter is connected to the carrier transmitter at the pumping station. The supervisory control equipment is arranged to be inoperative to the carrier impulses resulting from the operation of the telemetering transmitter, when the telemetering equipment is thus connected to the carrier channel. The indication of the amount of water being pumped may be retained at the water treatment plant as long as desired.

Release of the telemetering indication and restoration of the supervisory control equipment is effected by restoring the pull type selection key just as for the release of other functions. However, in this instance, keying of the carrier transmitter at the water treatment plant for the supervisory control reset impulse is automatically held off until completion of a

telemetering impulse, which may be in progress. The reset impulse is transmitted at the end of the telemetering impulse and operates the supervisory control receiver relay at both the water treatment plant and the pumping station. The duration of the supervisory control reset impulse is approximately one-fifth of a second so that the reset impulse can be completed in the off period of the telemetering transmitter even though the minimum off period of 0.5 second is in effect due to full scale reading being obtained.

The supervisory control receiver relay at the pumping station operates on all telemetering impulses when the telemetering function is selected. Operation of both the telemetering transmitter contact and the supervisory receiver relay does not allow the supervisory control equipment to reset. However, operation of the supervisory control receiver relay with the telemetering transmitter contact open, such as results when the reset impulse is transmitted from the water treatment plant, causes the equipment to reset and restore to normal, releasing the supervisory control equipment for other functions.

Control of Electric Motor and Gasoline Engine at Pumping Station

Figure 8 is a schematic diagram of the automatic and manual control for the 75-horsepower electric motor and the 180-horsepower gasoline engine which are used to pump water from the pumping station to the water treatment plant through the 24-inch pipe.

An automatic–manual transfer scheme which can be operated either from the water treatment plant or the pumping station by means of the supervisory control equipment is employed. A latching type relay 43 is used to effect this transfer. It is to be noted from the schematic diagram that the coils of this relay can be operated either from a push button station located at the pumping station or from supervisory control interposing relays.

When the manual contacts of the transfer relay 43 are closed, the electric motor and the gasoline engine are independently under the control of local push button stations and the supervisory control equipment. Thus, when on manual control, the operator at the water treatment plant can start or stop either the electric motor or the gasoline engine at will. Either unit can be operated to pump water or both units can be operated to pump water at the same time.

Manual Control of Electric Motor

When the electric motor is started on manual control from either the push button station at the pumping station or from the water treatment plant by means of the supervisory control equipment, motor control relay $4M$ is first energized. This relay seals itself in through a make contact of auxiliary voltage relay $27X$. Contacts of motor control relay $4M$ energize the solenoids of the normal and emergency cone values of the pump driven by the electric motor providing there is no surge of water at the time. When the limit switch on the normal cone valve closes, motor start relay 6 is energized. A contact of relay 6 completes the circuit to the motor breaker closing relay $52X$ and relay $52X$ in turn energizes the closing coil of the motor breaker. The motor thus is started and water is pumped.

If the electric motor is stopped by means of the push button at the pumping station, by the supervisory control equipment, or if voltage relay 27 closes its back contact due to loss of voltage or low voltage, motor control relay $4M$ is de-energized. This de-energizes the normal cone valve solenoid of the electric motor pump. When the limit switches on the normal cone valve open, the emergency cone valve solenoid is de-energized and motor start relay 6 is de-energized. A contact of relay 6 energizes the trip coil of the motor breaker, stopping the motor.

Emergency shutdown of the electric motor results if overcurrent relay 51 or thermal overload relay 49 operate. Operation of either of these relays energizes lockout relay 86 which must be manually reset. Relay 86 simultaneously de-energizes motor control relay $4M$ and trips the motor breaker.

Manual Control of Gasoline Engine

When the gasoline engine is started on manual control from either the push button station at the pumping station or from the water treatment plant by means of the supervisory control equipment, engine control relay $4E$ is first energized. This relay then seals itself in. Contacts of engine control relay $4E$ energize the solenoids of the normal and emergency cone valves of the pump driven by the gasoline engine and also complete a circuit to the gasoline engine starting panel, the detailed operation of which is not shown on the schematic diagram. Thus, the cone valves start to open and the gasoline engine starts providing there is no surge on the water line and the oil pressure of the engine is up.

When the gasoline engine is stopped from either the push button station at the pumping station or by the supervisory control equipment, engine control relay $4E$ is de-energized. When relay $4E$ is de-energized it immediately releases the solenoid of the normal cone valve of the pump driven by the gasoline engine.

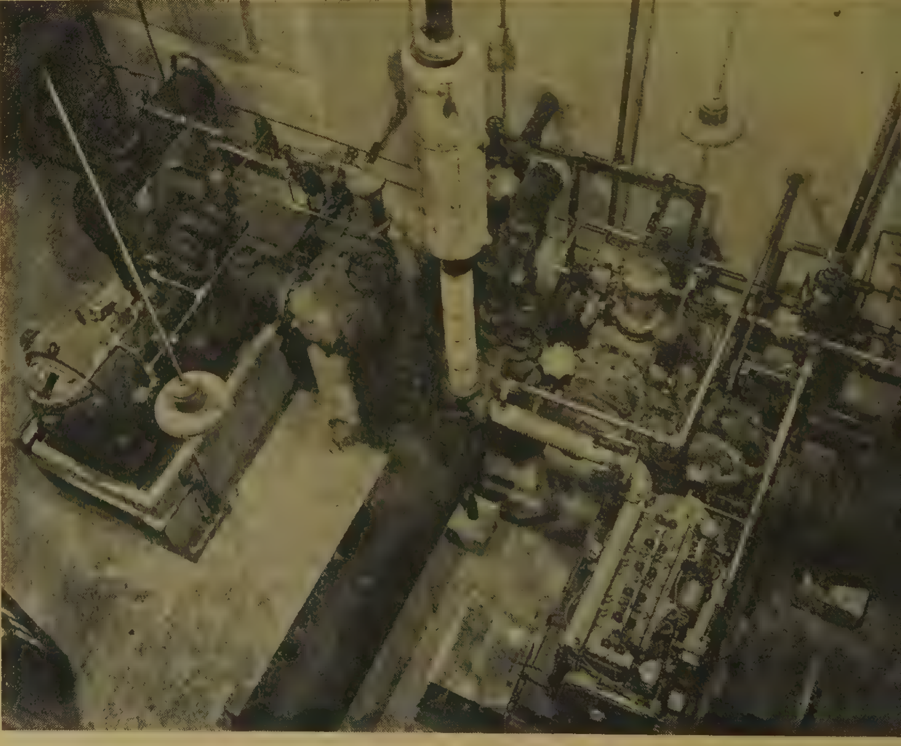


Figure 9. View showing electric motor and gasoline engine in pumping station

When the limit switch on the normal cone valve opens it releases the solenoid of the emergency cone valve and causes the engine control to transfer to the idling cycle.

AUTOMATIC CONTROL

With the automatic contacts of the transfer relay closed, the electric motor will be running if there is voltage available and the lockout relay is not operated. When on automatic control the electric motor cannot be stopped from either the push button station or from the water treatment plant by means of the supervisory control equipment. If a transfer from manual control to automatic control is effected from either the pumping station or the water treatment plant the electric motor will start if it was not already running and the gasoline engine will stop if it was running.

If when on automatic control the 4-kv voltage should fail at the pumping station or the lockout relay 86 should operate, the electric motor is stopped and the gasoline engine is started automatically. If the electric motor stops as a result of loss of voltage and voltage again becomes available the gasoline engine automatically stops and the electric motor starts.

When on automatic control, either the electric motor or the gasoline engine will be running if conditions are satisfactory for either to be running, with preference being given to the electric motor. The electric motor and the gasoline engine never will pump water simultaneously when on automatic control.

Conclusions

Power-line carrier equipment has been applied to transmission lines for more than 20 years. The majority of the applications have been for protective relaying and communication with the carrier in most instances being operated over open construction transmission lines rated 33 kv or greater. While there is still much information to be obtained concerning operation of power-line carrier equipment over such lines considerable empirical data have resulted from tests made on some of these applications so that the performance of the carrier equipment usually can be predicted fairly accurately for proposed applications.[2]

Up to the present time there have been very few applications of power-line carrier equipment to low voltage lines. This is especially true in the case of power cable. The possibility of combined functions on carrier channels makes it economically sound to apply carrier equipment to lower voltage lines in many instances, and it is expected that there will be many such applications in the future. The results obtained on the installation described indicate that standard carrier sets which will operate through a 30-decibel loss can be operated over considerable distances of power line cable. There is a need for similar installation data on future applications so that operation over cable can be predicted as accurately as operation over open construction power lines.

This installation of power-line carrier on a low voltage line indicates that there will be many applications of carrier equipment to low voltage lines (either open construction or cable) where it will be desirable to consider coupling means other than the standard line of coupling capacitors from an economic and space standpoint.

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Factors Affecting Insulation Resistance of Large D-C Machines

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INSULATION RESISTANCE of rotating machines was considered by many to be a haphazard collection of unrelated numbers. One reason for this view was that insulation resistance readings rarely were taken under the same conditions. Many of the factors which greatly affect insulation resistance were not known. In the past decade, significant contributions have been made toward a better understanding of the problem.[1–4] The influence of insulation temperature, time of application, and magnitude of test voltage, moisture, and machine size were recognized.

Study of the problem by the AIEE subcommittee on insulation resistance has resulted in the preparation of a proposed "Recommended Practice for Insulation Resistance Testing of A-C Rotating Machines." Its purpose is to summarize existing knowledge, make specific recommendations for test procedures, and give minimum insulation resistance formulas for new machines. Reference to d-c machines was omitted because it was recognized that the problem on these machines in some respects was different, and that further study was necessary. It was hoped, however, that in the near future sufficient progress would be made on the d-c problem to permit extension of the proposed recommended practice to include both classes of machines.

To assist in this problem, insulation resistance data on new d-c machines over 1,000 horsepower have been accumulated by the authors and their associates over a 5-year period. The purpose of the present work has been to determine, from a statistical examination of these data, the relative effect of the principal factors which affect the insulation resistance of d-c machines. In addition to previously recognized factors, such as machine size and insulation temperature, the absolute humidity of the surrounding air has been found to have an important effect on the insulation resistance of commutator type machines. This study has led to the development of an insulation resistance formula for d-c machines which includes all of these factors. It has been recognized

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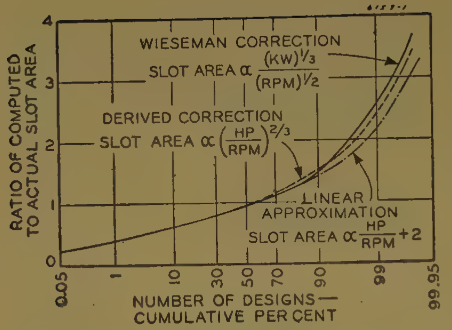

Figure 1. Frequency distribution of the ratio of computed to actual slot area showing correlation of various formulas

Data based on representative designs of 200 d-c machines of 100 horsepower and more

by the authors that for greatest usefulness, such a formula should express the relationship of the various factors in terms of simple arithmetical processes. Accordingly, a linear relationship involving rated horsepower, speed, and the winding insulation dimensions has been developed and compared to the existing formula. The correction factor for the effect of absolute humidity at the time of test, which has been included in the formula, is a new concept.

The factors which affect measured values of machine insulation resistance may be grouped as follows:

1. The nature and condition of the insulation material.
2. The design and construction of the machine.
3. The conditions of test.

Quantitative evaluation of the effect of these influences is difficult because of the complex structure of machine insulation. There is also a lack of basic knowledge of the fundamental processes involved and a lack of control of factors which have an important effect on machine insulation resistance.

Machine Insulation Structure

In general, insulation leakage between machine components may be segregated into two parallel paths, one through the body of the insulation itself, the other over the surface of the insulation. The relative magnitudes of the surface and volume leakage currents depend on the physical proportions of the insulation structure and the extent to which each is affected by existing conditions. It is usually difficult to determine the relative extent that each of these currents contributes to the total insulation leakage. In the case of stator windings of a-c synchronous and induction machines, it is perhaps reasonable to assume that surface leakage currents have a negligible effect on the total leakage current. Experience has shown, however, that at least under normal operating conditions the insulation resistance of d-c machines is determined largely by creepage factors. The same also appears to be true of new d-c machines under certain conditions of atmospheric humidity.

The basic difference between a-c synchronous and induction machines and d-c machines from an insulation standpoint is the existence of relatively large exposed creepage areas which are inherent in the design of d-c machines. This is caused principally by the commutator and to a lesser extent by the brush rigging, series, commutating, and compensating windings. Therefore, it is reasonable to expect that the insulation resistance characteristics of commutator type machines would be different from machines without commutators and that, in general, the commutator type machines would have lower values of insulation resistance.

Scope of Work

The insulation structure of large d-c machines usually consists of armature circuit insulation and shunt field insulation. The insulation resistance of the shunt field winding is usually relatively high. The armature circuit insulation resistance is of principal importance because it is usually lower and more affected by external conditions. The insulation of the armature circuit includes that of the armature winding, brush rigging, series, commutating, and compensating windings. The scope of the present work has been to study and analyze the results of armature circuit insulation resistance measurements of approximately 1,100 d-c machines in the range of 1,000 to 7,000 horsepower. Although insulation resistance data were not available for machines of less than 1,000 horsepower, it was felt by the authors that the dividing line between small and large d-c machines should be 100 horsepower. Therefore, in the development of the size correction which is included in the suggested minimum formula, machines of 100 horsepower and higher were considered.

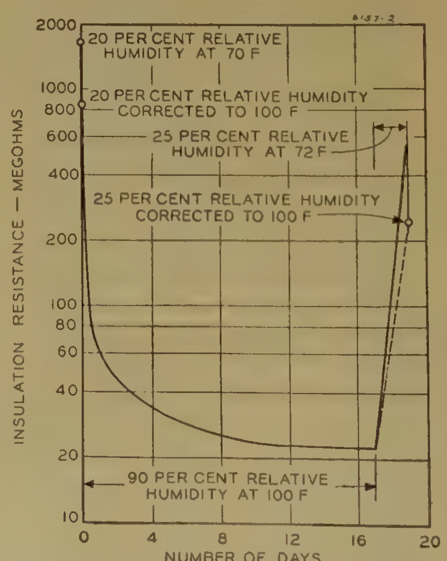

Figure 2. Insulation resistance decrease and recovery of a d-c armature exposed to 90 per cent relative humidity at 100 degrees Fahrenheit

In this work effort was directed toward estimating from a statistical study of the insulation resistance data known design proportions of the machines and known conditions of test, the probable influence of the principal factors which contribute to the measured value of insulation resistance. A further purpose of this work has been to relate these factors into a simple usable formula for expressing expected insulation resistance of new machines in terms of parameters which are known or easily obtainable. In the development of the ideas advanced in this paper, only general purpose d-c machines of ratings of 100 horsepower or more were considered. The following, therefore, does not necessarily apply to special types of d-c machines such as are used in the traction field.

Development of Formula Relating Rated Horsepower and Speed to Machine Size

One of the factors which influences measured values of insulation resistance, which is not related to the nature or condition of the insulation, is the physical size of the machine. A major shortcoming of the present AIEE minimum insulation resistance formula is that it makes no significant allowance for this factor. More rational formulas were proposed in 1934 by Wieseman which take into account effective insulation area and thickness. Although basically sound, these formulas have not enjoyed widespread use, largely because of a general unwillingness to handle formulas involving fractional powers. For greatest usefulness in a practical insulation resistance formula, a relationship among the various factors which involve relatively simple arithmetical processes is desirable. Accordingly, an attempt has been made in this work to develop a linear formula for armature

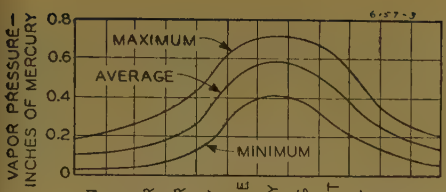

Figure 3. Record of the average and range of the daily vapor pressures for the Pittsburgh area for 1940 to 1946

circuit insulation resistance of d-c machines 100 horsepower and higher which includes the effect of machine size.

Under conditions where surface creepage is negligible, machine insulation resistance would be expected to vary inversely as the ratio of insulation contact area to thickness. Where surface creepage has an appreciable effect, the logic of making a size correction based on contact area, at first glance, would appear questionable. More careful analysis, however, reveals that there is a substantially linear correlation between contact area and effective creepage area. Therefore, even where surface creepage is not negligible, it appears that insulation contact area can be used as a basis for size correction.

The considerations involved and assumptions made in the development of a linear formula, relating machine size to rated voltage, horsepower, and speed are as follows:

1. The total insulation area under consideration includes that of all the windings in the armature circuit. In the development of the formula, only the insulation area in the armature slots is considered. The other areas involved are comparatively small and, in addition, are roughly proportional to the slot contact area.

2. The slot contact area of the armature is equal to the product of the slot perimeter, the number of slots, and the length of the armature iron.

3. Although slot insulation thickness is, in general, a function of voltage, the slot insulation thickness for large d-c machines is substantially constant. Therefore, rated voltage need not be included in the formula.

4. The machine output for a given temperature rise and average conditions of electric and magnetic loading is related to armature size by the following fundamental design relationship:

$$\frac{\text{Horsepower}}{\text{Revolutions per minute}}=K_1d^2l$$

where d and l are the armature diameter and length, respectively.

5. The slot depth is approximately constant.

6. An average value is assumed for the slot pitch (center to center distance between slots). Actually, this value may vary by a maximum of plus or minus 30 per cent.

7. An average value is assumed for the ratio of armature length to diameter. This value may vary by as much as plus or minus 80 per cent of the average value. Slot area, however, varies as the cube root of this ratio.

From the foregoing it follows that

$$\text{Slot area}=\frac{K_2dl}{\text{slot pitch}}=K_3dl=K_3ad^2$$

where $a=l/d$.

$$\frac{\text{Horsepower}}{\text{Revolutions per minute}}\propto d^2l\propto ad^3$$

$$d\propto\left[\frac{\text{horsepower}}{(\text{revolutions per minute})a}\right]^{1/3}$$

where d, l, and slot pitch are in inches, slot area is in square inches.

$$\text{Slot area}=K_4\left[\frac{\text{horsepower}}{(\text{revolutions per minute})}\right]^{2/3} \quad (1)$$

The best linear approximation for equation 1 was determined and is as follows:

$$\text{Slot area}=K_5\left(\frac{\text{horsepower}}{\text{revolutions per minute}}+2\right) \quad (2)$$

Actually, the values K_4 and K_5 (equations 1 and 2) are not constants because of the assumption of average values for design factors which affect machine size without directly affecting machine output. Average values, however, were obtained by substituting actual slot areas of 200 representative designs into the formula and solving for K_4 and K_5 in equations 1 and 2. The averages thus obtained for K_4 and K_5 were found to be 4,800 and 1,500, respectively.

The validity of the foregoing formula was checked by comparing the ratio of computed to actual slot areas for each formula. The computed ratios for the 200 representative designs were arranged in increasing order and plotted on arithmetic probability paper (see the appendix). Figure 1 shows a comparison of correlation for equations 1 and 2 as well as the Wieseman formula for 200 representative designs of d-c machines in the range of 100 to 7,000 horsepower. For purposes of comparison, a constant for the Wieseman formula was determined in the same manner as K_4 and K_5. It is apparent that the linear formula compares favorably with the other formulas and also has the advantage of simplicity. A size correction based on this linear relationship therefore is included in the minimum insulation resistance formula suggested by the authors.

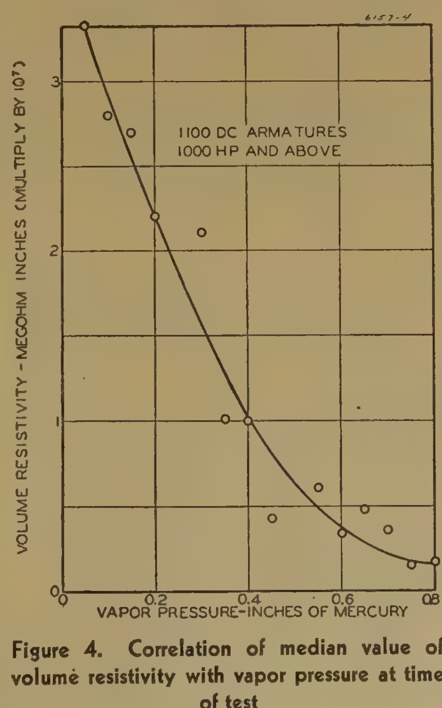

Figure 4. Correlation of median value of volume resistivity with vapor pressure at time of test

Effect of Absolute Humidity

Insulation resistance of all rotating machines is affected by prevailing atmospheric moisture conditions. For this reason, machines built and tested during the winter months, in general, have appreciably higher insulation resistance values than identical machines built and tested during the summer months. This effect is, in general, greater on commutator type machines where large creepage areas from bare copper to ground exist. Study of the data reveals that insulation resistance of the armature circuit of d-c machines is affected by a combination of atmospheric moisture conditions at the time of test, and during the period of manufacture.

The insulation resistance measurements of the armature circuit windings discussed in this work were made shortly after the running test. Since condensation of atmospheric moisture on the insulation surfaces greatly distorts insulation resistance, all tests were made with insulation temperatures five to ten degrees above room conditions. Measurements were made at 500 volts d-c and one minute after voltage application.

The effect of extreme humidity conditions on the insulation resistance of a d-c armature is illustrated in Figure 2. This armature which was initially at equilibrium at 70 degrees Fahrenheit and 20 per cent relative humidity, was exposed to 90 per cent relative humidity at 100

degrees Fahrenheit for 17 days and then returned to the initial ambient conditions. A reduction to 1/13 of the initial resistance after making corrections for temperature was observed at the end of the first day. The temperature correction is based on a ten to one decrease in insulation resistance for a 50-degree-centigrade temperature increase. A further reduction of one half was observed for the succeeding 16 days. Upon returning to room conditions of humidity and temperature, a rapid recovery to approximately one half of the initial insulation resistance value was observed. It, therefore, is apparent that the insulation resistance of a d-c armature is affected greatly by the prevailing atmospheric conditions at the time of test.

To evaluate the effect of atmospheric moisture conditions on insulation resistance, correlations of absolute and relative humidity at time of test with insulation resistivity were attempted. The insulation resistance of the 1,100 machines was corrected for machine size by multiplying the measured values by the ratio of the effective slot area and insulation thickness. The units of insulation resistivity thus obtained were megohm-inches.

Comparison of insulation resistivity with relative humidity at time of test revealed no significant correlation. However, since it was evident from Figure 2 that humidity conditions at time of test should affect the insulation resistance significantly, a correlation between insulation resistivity and absolute humidity was attempted. Absolute humidity differs from relative humidity in that it is a quantitative measure of atmospheric moisture. Relative humidity merely expresses the ratio of actual atmospheric moisture to the maximum or saturation conditions for a given temperature. Since absolute humidity at saturation varies logarithmically with temperature, it is evident that relative humidity by itself is not a measure of atmospheric moisture.

For practical purposes vapor pressure in inches of mercury is a convenient measure of absolute humidity. A record of daily average and range of vapor pressures in the Pittsburgh district for the years 1941 through 1945 is shown in Figure 3.* These observations were made less than five miles from the place of manufacture and were assumed to represent ambient moisture conditions at the plant.

Figure 4 shows the correlation of median insulation resistivity with vapor pressure at the time of test (see the appendix). Based on this correlation, all armature insulation resistivities were corrected to a vapor pressure of 0.1 inch of mercury. Figure 5 shows a comparison of the uncorrected insulation resistance data, the same data corrected for machine size, and the same data again corrected for vapor pressure at the time of test. It is evident from Figure 5 that corrections for machine size and vapor pressure based on the foregoing correlation produce an improvement in the central tendency of the data.

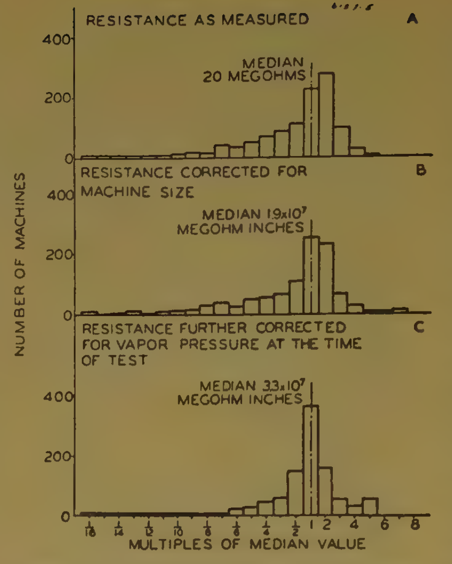

Figure 5. Comparison of insulation resistance frequency distributions before and after corrections—1,100 d-c machines

Although the correlation shown in Figure 4 is based on vapor pressure at time of test, it actually includes the effect of vapor pressure during manufacture. This is true because, in most cases, the average vapor pressure conditions during manufacture and test are substantially the same. For purposes of applying a correction for vapor pressure at the time of test in a minimum insulation resistance formula, it is desirable to separate the effect of vapor pressure during manufacture from that at the time of test.

Table I. Correlation of Average Resistivity of 700 Identical D-C Armatures by Months, With Average Vapor Pressure at Month of Test and Two Months Prior to Test

Month	Avg Vapor Pressure, In. Hg	Uncorrected Avg Resistivities, Megohm-In. $\times 10^7$	Avg Resistivities Corrected to 0.1 In. Hg, Megohm-In. $\times 10^7$
January	0.14	2.9	6.9
February	0.13	2.5	4.8
March	0.15	3.8	8.0
April	0.20	2.7	7.0
May	0.37	1.3	7.2
June	0.57	0.56	6.5
July	0.54	0.58	11.5
August	0.53	0.26	7.8
September	0.36	0.43	8.3
October	0.25	1.7	22.6
November	0.17	2.9	17.7
December	0.12	3.0	9.0
January	0.14	3.5	8.3
February	0.14	3.1	5.2

An examination was made of the insulation resistance of a large number of identical machines which were built at a rate of approximately 50 machines per month over a 15-month period. Table I shows the average insulation resistance of these armatures and the average monthly vapor pressures. If the following assumptions are made, it will be seen from column 3 of Table I that in most cases reasonable approximations for the effect of vapor pressure can be made:

1. Insulation resistance is inversely proportional to vapor pressure at time of test. Figure 2 shows a 13 to 1 initial reduction in insulation resistance for a 13 to 1 increase in vapor pressure.

2. Insulation resistance is inversely proportional to the vapor pressure at time of insulation varnish treatment. In the case of these machines, the time of varnish treatment is assumed to be two months prior to the month of test.

3. The over-all correction for manufacturing and test conditions is the product of the individual corrections for each influence. This assumption was made because it was found from inspection of the data of Table I that a correction of this sort correlated the observed data. Since only the creepage component of insulation resistance is affected by the vapor pressure at the time of test, the foregoing assumption implies that the principal effect of the humidity conditions during manufacture is on the creepage resistance.

Based on the preceding assumptions, a formula for correcting all the observed resistivities to constant vapor pressure conditions of 0.1 inch of mercury is as follows:

$$\text{Corrected resistivity} = \text{uncorrected resistivity} \times \frac{VP_1}{0.1} \times \frac{VP_2}{0.1}$$

where VP_1 and VP_2 are vapor pressures at the time of test and two months prior to the time of test, respectively. Column 3 of Table I indicates that, except for the months of October and November, monthly average resistivities corrected to constant vapor pressure conditions are approximately the same. Corrections, therefore, were made individually to the entire group of machines excluding those tested in October and November. Figure 6 gives a comparison of the frequency distribution of uncorrected and corrected data grouped according to multiples and

* Data obtained from original records of the Pittsburgh Weather Bureau.

common fractions of the median value of resistivity. Figure 7 shows a frequency distribution of the corrected data grouped in the usual manner. As shown in Figures 6 and 7, the corrections significantly improve the normality of the observed data. Therefore, the assumptions of the relative effect on insulation resistance of the ambient humidity conditions during manufacture and test appear to be reasonable.

After a machine is in operation, the effect of moisture conditions at the time of varnish treatment would be expected to diminish, and in time the moisture content of the insulation is determined by the prevailing ambient and operating conditions. The real criterion of the moistureproof quality of machine insulation is not necessarily its moisture content, but rather the rate at which moisture is absorbed under given conditions. Under conditions of operation, accumulations of conducting particles on the exposed creepage surface may have a predominating influence on insulation resistance.

Suggested Minimum Formula

The principal requirement for an ideal minimum insulation resistance formula is a simple relationship between the factors which are known to affect the machine insulation resistance. These factors should be in terms of parameters which are known or are easily obtainable at the time of test.

1. *Variation of slot area*, for d-c machines in the range of 100 to 7,000 horsepower, has a maximum range of 100 to 1. A simple relationship between output and speed rating has been developed which makes an average correction with a 27 per cent dispersion. Since slot area varies to a certain extent independently of the output or speed rating, significantly closer correction does not appear possible.

2. *Variation of the exposed creepage surface of the armature* is approximately linearly related to the slot area of the armature so that any corrections made for slot area also will apply to this factor.

3. *Insulation wall thickness*, because of the limited voltage range of d-c machines, may be considered constant.

4. *Insulation temperature* for new d-c machines, where insulation resistance decreases by a factor of approximately ten, increases from 25 to 75 degrees centigrade. There is a relatively small amount of data on temperature-insulation resistance slopes of d-c machines. More data are needed to establish average values and probable dispersions for machines both when new and under operating conditions. In the meanwhile the effect of temperature is minimized best by consistently measuring the insulation resistance at approximately the same temperature. Five to ten degrees above ambient would be a convenient temperature. The possibility of moisture condensation then would be eliminated. If measurements are made at or close to room condition, temperature correction to 25 degrees centigrade, based on an assumed average slope, will introduce relatively small errors.

5. *Absolute humidities immediately preceding the time of test* have been shown to affect the measured insulation resistance by as much as ten to one for variations normally encountered. A simple correction for this factor in terms of ambient temperature and relative humidity at the time of test is possible for this variable.

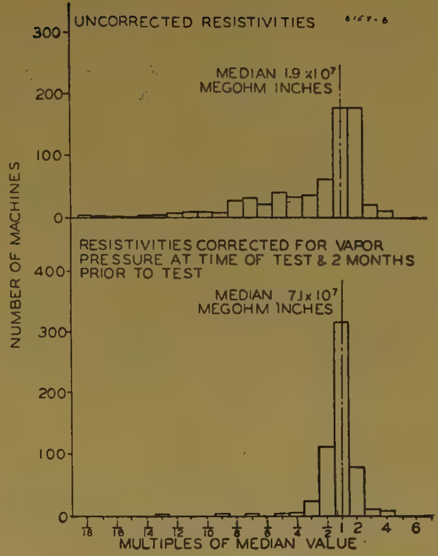

Figure 6. Comparison of insulation resistivity frequency distributions for 700 identical d-c machines before and after corrections for vapor pressure

It might be pointed out here that insulation temperature, as well as absolute humidity, has a diminishing effect on insulation resistance of machines operated in service under conditions where an accumulation of conducting dirt is likely to be deposited on the exposed creepage surfaces.

Assuming that a simple formula which gives the average expected insulation resistance of a d-c armature is possible, the selection of a suitable minimum insulation resistance formula depends upon what this formula thus determined is intended to represent. If the criterion is the value which is safe for application of the proof voltage test, the level of insulation resistivity set by the formula, if moisture is considered to be the contaminating agent, should depend primarily on when the insulation contains a sufficient quantity of moisture to affect adversely the short-time dielectric strength of the insulation. It does not appear possible at the present time to state definitely what this resistivity should be. Exploratory tests for paper-backed sheet-form mica insulation indicate that the short-time breakdown does not begin to be affected significantly until the resistivity is of the order of 1/10,000 or less of the average new machine insulation resistivity. However, while moisture contamination is the principal source of low insulation resistance of new machines, it is not the only possible cause. Values of insulation resistance, which would be considered safe if the assumption could be made that the cause was uniformly distributed moisture, might not be safe if caused by some other factor such as defective or damaged insulation. It should be emphasized that a high value of insulation resistance is no guarantee of adequate dielectric strength for test or operation. Insulation defects are possible which do not affect the insulation resistance. At best, insulation resistance indicates the existence and approximate extent of contaminating influences which affect insulation leakage. It appears that the best procedure possible at the present time in the selection of a minimum formula is to reach some agreement as to

1. What the average resistivity of new machine insulation should be.

2. Some reasonable method by which the minimum insulation resistance shall be calculated from this average value. It should be recognized that the selection of this minimum has no logical basis except that it represents the best estimate possible at the present time.

Table II. Correction Factor for Absolute Humidity at Time of Test in Terms of Ambient Temperature and Relative Humidity

Ambient Temperature, Deg F	Correction Factor*
10	0.18
20	0.30
30	0.48
40	0.70
50	1.00
60	1.47
70	2.15
80	2.95
90	3.90
100	5.50

* f_h = factor × per cent relative humidity.

Based on the foregoing considerations, the following formula is suggested for minimum insulation resistance of new d-c armatures of ratings of 100 horsepower and higher. Essentially, the formula is an expression for the expected average armature circuit insulation resistance of a new machine after correction for machine size, insulation temperature at the time of test, and absolute humidity at the time of test. Since the ground insulation on

d-c machines is essentially the same in both cases, no distinction is made between class A and class B insulation. Also, for general purpose d-c machines, the voltage range is limited and insulation wall thickness is substantially unaffected by rated voltage. Therefore, the voltage factor has been omitted in the formula.

$$R_t = \frac{4{,}000}{\left(\frac{\text{horsepower}}{\text{revolutions per minute}} + 2\right) f_h}$$

R_t = minimum insulation resistance in megohms at 25 degrees centigrade

f_h = correction factor for absolute humidity at time of test (see Table II)

In addition to this formula, a 2-megohm minimum insulation resistance is suggested for armatures when the calculated value from the formula is below this value. In general, these armatures would be large and important, and there is the ever-present possibility that the low insulation resistance might not be caused by moisture contamination, but might be the result of other localized defects.

Summary and Conclusions

Analysis of insulation resistance data of 1,100 d-c machines in the range of 1,000 to 7,000 horsepower has resulted in the rationalization of the complex factors involved. Consideration and segregation of the principal factors which influence measured values of insulation resistance has enabled a better understanding of the relative effect of each. In addition to previously recognized factors such as material resistivity, insulation temperature, and machine size, the absolute humidity of the surrounding air has been found to affect significantly the insulation resistance of commutator type machines. The total effect of humidity on new machines at the factory has been shown to consist of

1. The effect of prevailing atmospheric humidity during insulation treatment.

2. The effect of atmospheric humidity at the time of test.

Prevailing atmospheric humidity during insulation treatment has only a transitory effect on the machine insulation resistance. Ultimate moisture concentration in machine insulation depends on service conditions. For commutator type machines, absolute humidity at the time of test is a test condition in the same sense

Figure 7. Frequency distribution of volume resistivity of 700 identical machines after correction for vapor pressure at time of test and two months prior to test

as insulation temperature. This effect would be expected to apply also to machines under service conditions, except where accumulation of conducting particles on exposed creepage surfaces has a predominating influence.

The correlation of these various factors has resulted in the development of a simple linear insulation resistance formula which is offered for consideration by the industry.

$$R_t = \frac{4{,}000}{\left(\frac{\text{horsepower}}{\text{revolutions per minute}} + 2\right) f_h}$$

where

R_t = minimum insulation resistance at 25 degrees centigrade for new d-c machines over 100 horsepower

f_h = correction factor for absolute humidity at the time of test

Appendix

Correlation of Computed Slot Area by Various Formulas and Actual Slot Area of Representative Designs

In the comparison of the correlation of the several formulas discussed in the text, use was made of graph paper having linear ordinate units and abscissa units arranged in an arithmetical probability scale. The ratio of computed to actual values for each formula was arranged in increasing order and expressed as a cumulative per cent of the total number of designs.

The interpretation of the ratio 1 corresponding to 50 per cent is that one half of the total number of designs have ratios of 1 or less. It is seen, therefore, that for the linear approximation the central 90 per cent of the distribution have ratios between 0.5 to 1.5. The scale of the abscissa is such that normal distributions plot as a straight line. The slope of this line is a measure of the dispersion and a smaller slope indicates smaller dispersion. Per cent dispersion, which for normal distributions is equal to $\frac{\text{standard deviation}}{\text{average}} \times 100$, may be obtained directly from the curve. The standard deviation includes approximately the central two thirds of the area distribution. Therefore, from the curve

$$\text{per cent dispersion} = \frac{(R_{83\cdot7} - R_{50}) \text{ or } (R_{50} - R_{16\cdot3})}{R_{50}} \times 100$$

where $R_{83\cdot7}$, R_{50}, and $R_{16\cdot3}$ are the ratios for the corresponding cumulative percentages in the distribution.

Correlation of Insulation Resistivity and Vapor Pressure at Time of Test

In this correlation the individual values of resistivity were classified according to vapor pressure at the time of test. For each vapor pressure group the median value of insulation resistivity was obtained. The median is defined as the value of the central unit of a distribution. If a distribution contains 101 numbers arranged in order of increasing value, the value of number 51 is the median of the distribution. The median was used as a basis for comparison because it gives less weight to extreme values of a distribution than does the arithmetic average.

Preparation of Table II

Table II lists absolute humidity corrections to be applied in the minimum formula. Absolute humidity is a function of the product of ambient temperature and relative humidity. Since ambient temperature and relative humidities are usually either known or easily determined, the absolute humidity correction is expressed in these parameters.

The table was constructed by considering as unity an average vapor pressure of 0.35 inch of mercury. This corresponds to a temperature and relative humidity of 50 degrees Fahrenheit and 100 per cent. Relative corrections for other temperatures and humidities were obtained from psychometric tables and expressed in terms of the assumed average conditions. In the proposed minimum formula, the average resistivity used is that corresponding to 0.35 inch of mercury.

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Properties and Uses of Thermistors—Thermally Sensitive Resistors

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Synopsis: A new circuit element and control device, the thermistor, or thermally sensitive resistor, is made of solid semiconducting materials whose resistance decreases about four per cent per degree centigrade. The thermistor presents interesting opportunities to the designer and engineer in many fields of technology for accomplishing tasks more simply, economically, and better than with available devices. Part I discusses the conduction mechanism in semiconductors and the criteria for usefulness of circuit elements made from them. The fundamental physical properties of thermistors, their construction, their static and dynamic characteristics, and general principles of operation are treated. Part II of this paper deals with the applications of thermistors. These include: sensitive thermometers and temperature control elements, simple temperature compensators, ultrahigh frequency power meters, automatic gain controls for transmission systems, voltage regulators, speech volume limiters, compressors and expandors, gas pressure gauges and flowmeters, meters for thermal conductivity determination of liquids, and contactless time delay devices. Thermistors with short time constants have been used as sensitive bolometers, and show promise as simple, compact, audiofrequency oscillators, modulators, and amplifiers.

I—Properties of Thermistors

THERMISTORS, or *therm*ally sensitive res*istors*, are devices made of solids whose electrical resistance varies rapidly with temperature. Even though they are only about 15 years old they already have found important and large scale uses in the telephone plant and in military equipments. Some of these uses are as time delay devices, protective devices, voltage regulators, regulators in carrier systems, speech volume limiters, test equipment for ultrahigh frequency power, and detecting elements for very small radiant power. In all these applications thermistors were chosen because they are simple, small, rugged, have a long life, and require little maintenance. Because of these and other desirable properties, thermistors promise to become new circuit elements which will be used extensively in the fields of communications, radio, electrical and thermal instrumentation, research in physics, chemistry and biology, and war technology. Specific types of uses which will be discussed in the second part of this paper include

1. Simple, sensitive, and fast responding thermometers, temperature compensators, and temperature control devices.
2. Special switching devices without moving contacts.
3. Regulators or volume limiters.
4. Pressure gauges, flowmeters, and simple meters for measuring thermal conductivity in liquids and gases.
5. Time delay and surge suppressors.
6. Special oscillators, modulators, and amplifiers for relatively low frequencies.

Before these uses are discussed in detail, it is desirable to present the physical principles which determine the properties of thermistors.

The question naturally arises: Why have devices of this kind come into use only recently? The answer is that thermistors are made of semiconductors and that the resistance of these can vary by factors up to a thousand or a million with surprisingly small amounts of certain impurities, with heat treatment, methods of making contact, and with the treatment during life or use. Consequently the potential application of semiconductors was discouraged by experiences such as the following: two or more units made by what appeared to be the same process would show large variations in their properties. Even the same unit might change its resistance by factors of two to ten by exposure to moderate temperatures or to the passage of current. Before semiconductors could be considered seriously in industrial applications, it was necessary to devote a large amount of research and development effort to a study of the nature of the conductivity in semiconductors, and of the effect of impurities and heat treatment on this conductivity, and to methods of making reliable and permanent contacts to semiconductors. Even though Faraday discovered that the resistance of silver sulphide changed rapidly with temperature, and even though thousands of other semiconductors have been found to have large negative temperature coefficients of resistance, it has taken about a century of effort in physics and chemistry to give the engineering profession this new tool which may have an influence similar to that of the vacuum tube, and may replace vacuum tubes in many instances.

If thermistors are to be generally useful in industry:

1. It should be possible to reproduce units having the same characteristics.
2. It should be possible to maintain constant characteristics during use, the contact should be permanent, and the unit should be chemically inert.
3. The units should be mechanically rugged.
4. The technique should be such that the material can be formed into various shapes and sizes.
5. It should be possible to cover a wide range of resistance, temperature coefficient, and power dissipation.

Thermistors might be made by any method by which a semiconductor could

Figure 1. Thermistors made in the form of a bead, rod, disk, washer, and flakes (left to right)

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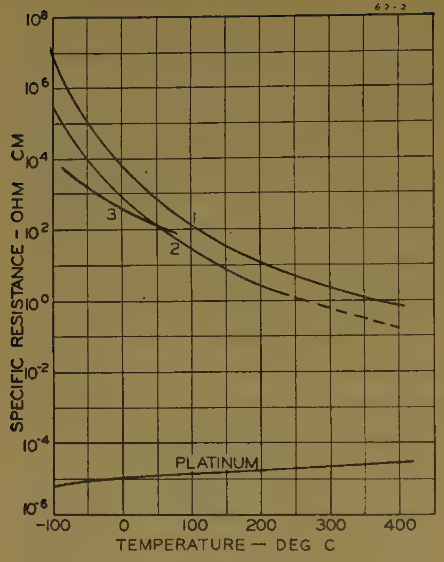

Figure 2. Logarithm of specific resistance versus temperature for three thermistor materials as compared with platinum

be shaped to definite dimensions, and contacts applied. These methods include

1. Melting the semiconductor, cooling, solidifying, cutting to size and shape.

2. Evaporation.

3. Heating compressed powders of semiconductors to a temperature at which they sinter into a strong compact mass, and firing on metal powder contacts.

Although all three processes have been used, the third method has been found to be useful most generally for mass production. This method is similar to that employed in ceramics or in powder metallurgy. At the sintering temperatures, the powders recrystallize and the dimensions shrink by controlled amounts. The powder process makes it possible to mix two or more semiconducting oxides in varying proportions, and obtain a homogeneous and uniform solid. It is thus possible to cover a considerable range of specific resistance and temperature coefficient of resistance with the same system of oxides. By means of the powder process, it is possible to make thermistors of a great variety of shapes and sizes to cover a large range of resistances and power handling capacities.

Figure 1 shows thermistors made in the form of beads, rods, disks, washers, and flakes. Beads are made by stringing two platinum alloy wires parallel to each other with a spacing of five to ten times the wire diameter. A mass of a slurry of mixed oxides is applied to the wires. Surface tension draws this mass into the form of a bead. From 10 to 20 such beads are evenly spaced along the wires. The beads are allowed to dry, and are heated slightly until they have sufficient strength so that the string can be handled. They then are passed through the sintering furnace. The oxides shrink onto the platinum alloy wires and make an intimate and permanent electrical contact. The wires then are cut to separate the individual beads. The diameters of the beads range from 0.015 to 0.15 centimeter with wire diameter ranging from 0.0025 to 0.015 centimeter.

Rod thermistors are made by mixing the oxides with an organic binder and solvent, extruding the mixture through a die, drying, cutting to length, heating to drive out the binder, and sintering at a high temperature. Contacts are applied by coating the ends with silver, gold, or platinum paste such as is used in the ceramic art, and heating or curing the paste at a suitable temperature. The diameter of the rods ordinarily can be varied from 0.080 to 0.64 centimeter. The length can vary from 0.15 to 5 centimeters.

Disks and washers are made in a similar way by pressing the bonded powders in a die. Possible disk diameters are 0.15 to 3 or 5 centimeters, thicknesses from 0.080 to 0.64 centimeter.

Flakes are made by mixing the oxides with a suitable binder and solvent to a creamy consistency, spreading a film on a smooth glass surface, allowing the film to dry, removing the film, cutting it into flakes of the desired size and shape, and firing the flakes at the sintering temperatures on smooth ceramic surfaces. Contacts are applied as described above. Possible dimensions are: thickness, 0.001 to 0.004 centimeter; length, 0.1 to 1.0 centimeter; width, 0.02 to 0.1 centimeter.

In any of these forms, lead wires can be attached to the contacts by soldering or by firing heavy metal pastes. The dimensional limits given above are those which have been found to be readily attainable.

In the design of a thermistor for a specific application, the following characteristics should be considered:

1. Mechanical dimensions including those of the supports.

2. The material from which it is made and its properties. These include the specific resistance and how it varies with temperature, the specific heat, density, and expansion coefficient.

3. The dissipation constant and power sensitivity. The dissipation constant is the watts that are dissipated in the thermistor, divided by its temperature rise in degrees centigrade above its surroundings. The power sensitivity is the watts dissipated to reduce the resistance by one per cent. These constants are determined by the area and nature of the surface, the surrounding medium, and the thermal conductivity of the supports.

4. The heat capacity, which is determined by specific heat, dimensions, and density.

5. The time constant. This determines how rapidly the thermistor will heat or cool. If a thermistor is heated above its surroundings and then allowed to cool, its temperature will decrease rapidly at first, and then more slowly until it finally reaches ambient temperature. The time constant is the time required for the temperature to fall 63 per cent of the way toward ambient temperature. The time constant in seconds is equal to the heat capacity in joules per degree centigrade, divided by the dissipation constant in watts per degree centigrade.

6. The maximum permissible power that can be dissipated consistent with good stability and long life, for continuous operation, and for surges. This can be computed from the dissipation constant and the maximum permissible temperature rise. This and the resistance-temperature relation determine the maximum decrease in resistance.

Properties of Semiconductors

As most thermistors are made of semiconductors, it is important to discuss the properties of the latter. A semiconductor may be defined as a substance whose electrical conductivity at, or near, room temperature is much less than that of typical metals, but much greater than that of typical insulators. While no sharp boundaries exist between these classes of conductors, one might say that semiconductors have specific resistances at room

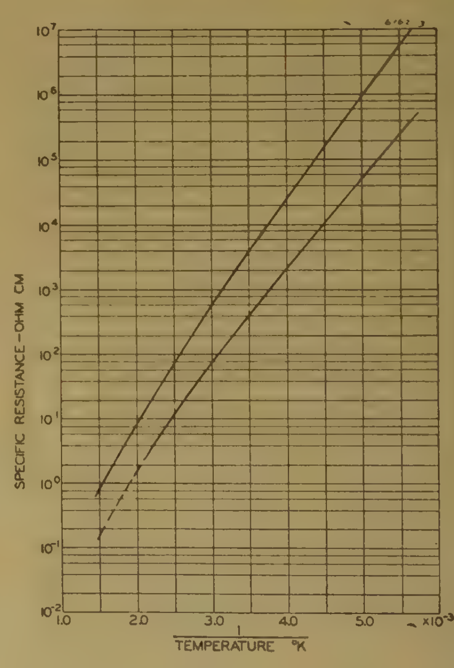

Figure 3. Logarithm of the specific resistance of two thermistor materials as a function of inverse absolute temperature

See equation 1

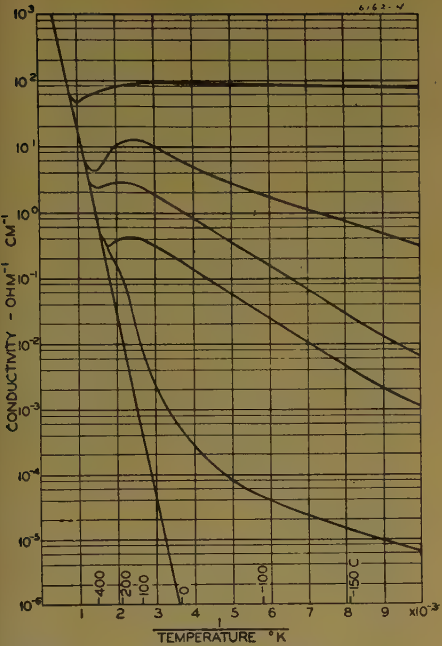

Figure 4. Logarithm of the conductivity of various specimens of silicon as a function of inverse absolute temperature

The conductivity increases with the amount of impurity

temperature from 0.1 to 10^9 ohm-centimeters. Semiconductors usually have high negative temperature coefficients of resistance. As the temperature is increased from 0 degrees centigrade to 300 degrees centigrade, the resistance may decrease by a factor of a thousand. Over this same temperature range the resistance of a typical metal, such as platinum, will increase by a factor of two. Figure 2 shows how the logarithm of the specific resistance, ρ, varies with temperature, T, in degrees centigrade for three typical semiconductors and for platinum. Curves 1 and 2 are for "material 1" and "material 2" which have been used extensively to date. "Material 1" is composed of manganese and nickel oxides. "Material 2" is composed of oxides of manganese, nickel, and cobalt. The dashed part of curve 2 covers a region in which the resistance-temperature relation is not known as accurately as it is at lower temperatures. Curve 3 is an experimental curve for a mixture of iron and zinc oxides in the proportions to form zinc ferrite. From Figure 2 it is obvious that neither the resistance R nor log R varies linearly with T.

Figure 3 shows plots of log ρ versus $1/T$ for "material 1" and "material 2". These do form approximate straight lines. Hence:

$$\rho=\rho_\infty \epsilon^{B/T} \text{ or } \rho=\rho_0\epsilon^{(B/T-B/T_0)} \qquad (1)$$

where T = temperature in degrees Kelvin, $\rho_\infty = \rho$ when $T = \infty$ or $1/T = 0$, $\rho_0 = \rho$ when $T = T_0$, ϵ = Naperian base = 2.718, and B is a constant equal to 2.303 times the slope of the straight lines in Figure 3. The dimensions of B are degrees Kelvin or degrees centigrade. B plays the same role in equation 1 as does the work function in Richardson's equation for thermionic emission. For material 1, B = 3,920 degrees centigrade. This corresponds to an electron energy of 3,920/11,600 or 0.34 volt.

While the curves in Figure 3 are approximately straight, a more careful investigation shows that the slope increases linearly as the temperature increases. From this it follows that a more precise expression for ρ is:

$$\rho=AT^{-c}\epsilon^{D/T}$$

or

$$\log \rho=\log A-c\log T+D/2.303T \qquad (2)$$

The constant c is a small positive or negative number or zero. For "material 1", log A = 5.563, c = 2.73, and D = 3,100. For a particular form of "material 2", log A = 11.514, c = 4.83, and D = 2,064.

If we define temperature coefficient of resistance, α, by the equation

$$\alpha=(1/R)(dR/dT) \qquad (3)$$

it follows from equation 1 that

$$\alpha=-B/T^2 \qquad (4)$$

For "material 1", and T = 300 degrees Kelvin, α = −3,920/90,000 = −0.044. For platinum, α = +0.0037, or roughly ten times smaller than for semiconductors, and of the opposite sign. From equation 2 it follows that

$$\alpha=-(D/T^2)-(c/T) \qquad (5)$$

From equation 3 it follows that

$$\alpha=(1/2.303)\ (d\log R/dT) \qquad (6)$$

For a discussion of the nature of the conductivity in semiconductors, it is simpler and more convenient to consider the conductivity, σ, rather than the resistivity, ρ.

$$\sigma=1/\rho \text{ and } \log\sigma=-\log\rho \qquad (7)$$

The characteristics of semiconductors are brought out more clearly if the conductivity or its logarithm are plotted as a function of $1/T$ over a wide temperature range. Figure 4 is such a plot for a number of silicon samples containing increasing amounts of impurity. At high temperatures, all the samples have nearly the same conductivity. This is called the intrinsic conductivity because it seems to be an intrinsic property of silicon. At low temperatures the conductivity of different samples varies by large factors. In this region silicon is said to be an impurity semiconductor. For extremely pure silicon, only intrinsic conductivity is present and the resistivity obeys equation 1. As the concentration of a particular impurity increases, the conductivity increases, and the impurity conductivity predominates to higher temperatures. Some impurities are much more effective in increasing the conductivity than others. One hundred parts per million of some impurities may increase the conductivity of pure silicon at room temperature by a factor of 10^7. Other impurities may be present in 10,000 parts per million and have a small effect on the conductivity. Two samples may contain the same concentration of an impurity and still differ greatly in their low temperature conductivity. If the impurity is in solid solution, that is, atomically dispersed, the effect is great; if the impurity is segregated in atomically large particles, the effect is small. Since heat treatments affect the dispersion of impurities in solids, the conductivity of semiconductors frequently may be altered radically by heat treatment. Some other semiconductors are not affected greatly by heat treatment.

The impurity need not be a foreign element; in the case of oxides or sulphides, it can be an excess or a deficiency of oxygen or sulphur from the exact

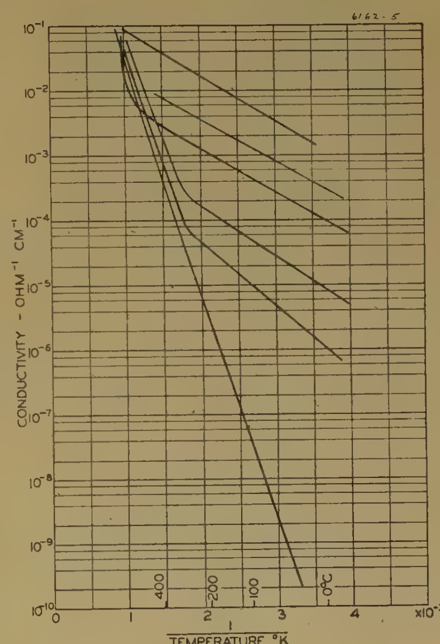

Figure 5. Logarithm of the conductivity of various specimens of cuprous oxide as a function of inverse absolute temperature

The conductivity increases with the amount of excess oxygen above the stoichiometric value in Cu_2O

Data from reference 1

stoichiometric relation. This excess or deficiency can be brought about by heat treatment. Figure 5 shows how the conductivity depends on temperature for a number of samples of cuprous oxide, Cu_2O, heat-treated in such a way as to result in varying amounts of excess oxygen from zero to about one per cent.[1] The greater the amount of excess oxygen, the greater is the conductivity in the low temperature range. At high temperatures, all samples have about the same conductivity.

Semiconductors can be classified on the basis of the carriers of the current into ionic, electronic, and mixed conductors. Chlorides such as NaCl and some sulphides are ionic semiconductors; other sulphides and a few oxides, such as uranium oxide, are mixed semiconductors; electronic semiconductors include most oxides such as Mn_2O_3, Fe_2O_3, NiO, carbides such as silicon carbide, and elements such as boron, silicon, germanium, and tellurium. In ionic and mixed conductors, ions are transported through the solid. This changes the density of carriers in various regions, and thus changes the conductivity. Because this is undesirable, they rarely are used in making thermistors, and hence we will concentrate our interest on electronic semiconductors.

The theoretical and experimental physicists have established that there are two types of electronic semiconductors which can be called N and P type, depending upon whether the carriers are negative electrons or are equivalent to positive "holes" in the filled energy band. In N type, the carriers are deflected by a magnetic field as negatively charged particles would be, and conversely for P type. The direction of deflections is ascertained by measurement of the sign of the Hall effect. The direction of the thermoelectric effect also fixes the sign of the carriers. By determining the resistivity, Hall coefficient, and thermoelectric power of a particular specimen at a particular temperature, it is possible to determine the density of carriers, whether they are negative or positive, and their mobility or mean free path. The mobility is the mean drift velocity in a field of one volt per centimeter.

The existence of these classifications is explained by the theoretical physicist[4-2] in terms of the diagrams in Figure 6. In an intrinsic semiconductor, at low temperatures the valence electrons completely fill all the allowable energy states. According to the exclusion principle, only one electron can occupy a particular energy state in any system. In semiconductors and insulators there exists a region of energy values, just above the allowed band, which are not allowed. The height of this unallowed band is expressed in equivalent electron volts, ΔE. Above this unallowed band there exists an allowed band, but at low temperatures there are no electrons in this band. When a field is applied across such a semiconductor, no electron can be accelerated, because if it were accelerated, its energy would be increased to an energy state which is either filled or unallowed. As the temperature is raised, some electrons acquire sufficient energy to be raised across the unallowed band into the upper allowed band. These electrons can be accelerated into a slightly higher energy state by the applied field and thus can carry current. For every electron that is put into an activated state there is left behind a hole in the normally filled band. Other electrons having slightly lower energies can be accelerated into these holes by the applied field. The physicist has shown that these holes act toward the applied field as if they were particles having a charge equal to that of an electron but of opposite sign, and a mass equal to, or somewhat larger than, the electronic mass. In an intrinsic semiconductor, about half the conductivity is due to electrons and half due to holes.

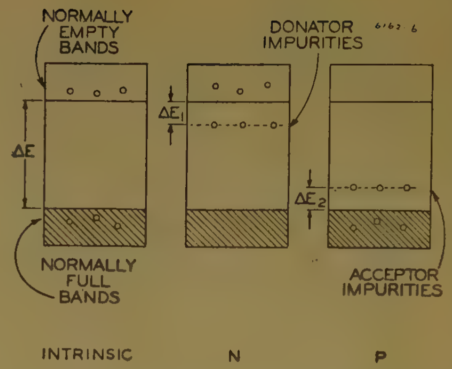

Figure 6. Schematic energy level diagrams illustrating intrinsic, *N*, and *P* types of semiconductors

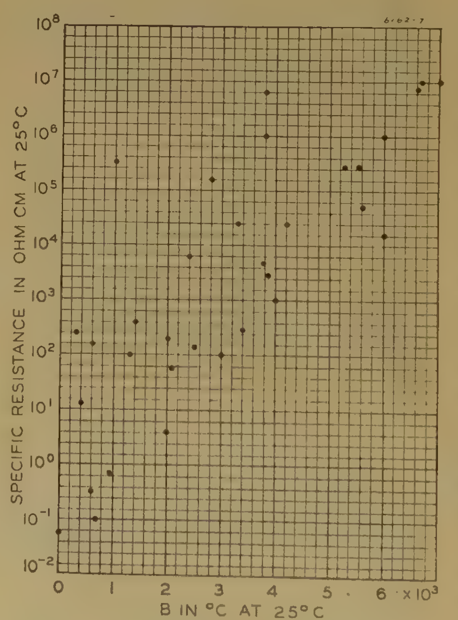

Figure 7. Logarithm of the resistivity of various semiconducting materials as a function of *B* in equation 1

The quantity, *B*, is proportional to the temperature coefficient of resistance as given in equation 4

The quantity ΔE is related to B in equation 1 by

$$2B = (\Delta E)e/k \tag{8}$$

in which B is in degrees centigrade, ΔE is in volts, e is the electronic charge in coulombs, k is Boltzmann's constant in joules per degree centigrade. The value of e/k is 11,600 so that

$$\Delta E = B/5{,}800 \tag{8a}$$

The difference between metals, semiconductors, and insulators results from the value of ΔE. For metals ΔE is zero or very small. For semiconductors ΔE is greater than about 0.1 volt, but less than about 1.5 volts. For insulators ΔE is greater than about 1.5 volts.

Some impurities with positive valencies which may be present in the semiconductor may have energy states such that ΔE_1 volts equivalent energy can raise the valence electron of the impurity atom into the allowed conduction band. See Figure 6. The electron now can take part in conduction; the donator impurity is a positive ion which usually is bound to a particular location and can take no part in the conductivity. These are excess or N type conductors. The conductivity depends on the density of donators, ΔE_1, and T.

Similarly, some other impurity with negative valencies may have an energy state ΔE_2 volts above the top of the filled band. At room temperature or higher, an electron in the filled band may be raised in energy and accepted by the impurity, which then becomes a negative ion and usually is immobile. However, the hole which results can take part in the conductivity.

In all cases represented in Figure 6, an electron occupying a higher energy level than a positive ion or a hole, has a certain probability that in any short interval of time it will drop into a lower energy state. However, during this same time interval there will be electrons which will be

raised to a higher energy level by thermal agitation. When the number of electrons per second which are being elevated is equal to the number which are descending in energy, equilibrium prevails. The conductivity, σ, is then

$$\sigma = Nev_1 + Pev_2 \tag{9}$$

where N and P are the concentrations of electrons and holes respectively, e is the charge on the electron, v_1 and v_2 are the mobilities of electrons and holes respectively.

This explains the following experimental facts which otherwise are difficult to interpret:

1. N-type oxides, such as ZnO, when heated in a neutral or slightly reducing atmosphere become good conductors, presumably because they contain excess zinc which can donate electrons. If they then are heated in atmospheres which are increasingly more oxidizing, their conductivity decreases until eventually they are intrinsic semiconductors or insulators.

2. P-type oxides, such as NiO, when heat-treated in strongly oxidizing atmospheres are good conductors. Very likely they contain oxygen in excess of the stoichiometric relation and this oxygen accepts additional electrons. When these are heated in less oxidizing or neutral atmospheres they become poorer conductors, semiconductors, or insulators.

3. When a P-type oxide is sintered with another P-type oxide, the conductivity increases; similarly for two N-type oxides. But when a P type is added to an N type the conductivity decreases.

4. If a metal forms several oxides, the one in which the metal exerts its highest valence is N type, while the one in which it exerts its lowest valence will be P type.[5]

For several reasons it is desirable to survey the whole field of semiconductors for resistivity and temperature coefficient. One way in which this might be done is to draw a line in Figure 3 for each specimen. Before long, such a figure would consist of such a maze of intersecting lines that it would be difficult to single out and follow any one line. The information can be condensed by plotting log ρ_0 versus B in equation 1 for each specimen.[6] The most important characteristics of a specimen thus are represented by a single point and many more specimens can be surveyed in a single diagram. Figure 7 shows such a plot for a large number of semiconductors investigated at these laboratories or reported in the literature. Values for ρ_0 and B are given for T = 25 degrees centigrade. The points form a sort of "milky way." Semiconductors having a high ρ_0 are likely to have a high value of B and vice versa. If a series of semiconductors has points in Figure 7 which fall on a straight line with a slope of 1/2.3 T_0, they have a common intercept in Figure 3 for $(1/T) = 0$.

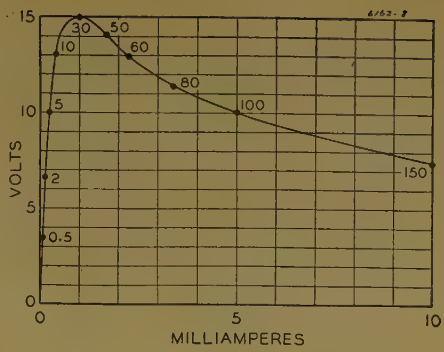

Figure 8. Static voltage–current curve for a typical thermistor

The numbers on the curve are the degrees centigrade rise in temperature above ambient temperature

Physical Properties of Thermistors

One of the most interesting and useful properties of a thermistor is the way in which the voltage, V, across it changes as the current, I, through it increases. Figure 8 shows this relationship for a 0.061-centimeter diameter bead of "material 1" suspended in air. Each time the current is changed, sufficient time is allowed for the voltage to attain a new steady value. Hence this curve is called the steady state curve. For sufficiently small currents, the power dissipated is too small to appreciably heat the thermistor, and Ohm's law is followed. However, as the current assumes larger values, the power dissipated increases, the temperature rises above ambient temperature, the resistance decreases, and hence the voltage is less than it would have been had the resistance remained constant. At some current, I_m, the voltage attains a maximum or peak value, V_m. Beyond this point, as the current increases the voltage decreases, and the thermistor is said to have a negative resistance whose value is dV/dI. The numbers on the curve give the rise in temperature above ambient temperature in degrees centigrade.

Because currents and voltages for different thermistors cover such a large range of values, it has been found convenient to plot log V versus log I. Figure 9 shows such a plot for the same data as in Figure 8. For various points on the curve, the temperature rise above ambient temperature is given. In a log log plot, a line with a slope of 45 degrees represents a constant resistance, and a line with a slope of −45 degrees represents constant power.

For a particular thermistor, the position of the log V versus log I plot is shifted, as shown in Figure 10, by changing the dissipation constant C. This can be done by changing the air pressure surrounding the bead, changing the medium, or changing the degree of thermal coupling between the thermistor and its surroundings. The value of C for a particular thermistor in given surroundings readily can be determined from the V versus I curve in either Figures 8 or 9. For each point, V/I is the resistance, while V times I is W, the watts dissipated. The resistance data are converted to temperature from R versus T, given by equation 2. A plot is then made of W versus T. For thermistors in which most of the heat is conducted away, W will increase linearly with T, so that C is constant. For thermistors suspended by fine wires in a vacuum, W will increase more rapidly than proportional to T, and C will increase with T. For thermistors of ordinary size and shape, in still air, C/area is equal to from 1 to 40 milliwatts per degree centigrade per square centimeter, depending upon the size and shape factor.

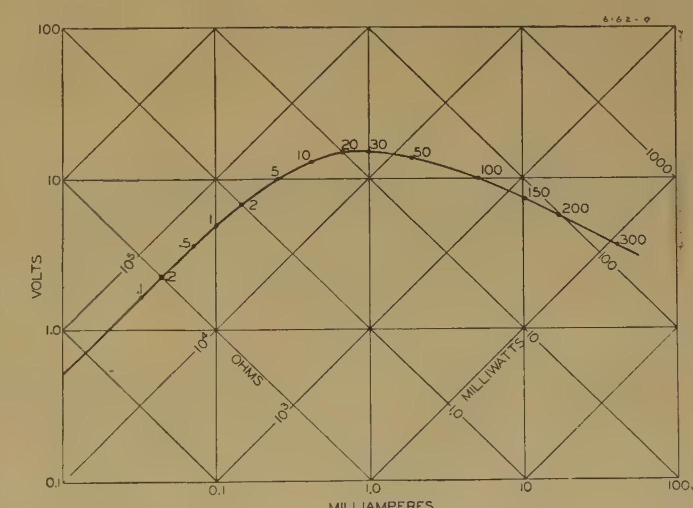

Figure 9. Logarithmic plot of static voltage–current curve for the same data as in Figure 8

The diagonal lines give the values of resistance and power

The user of a thermistor may want to know how many watts can be dissipated before the resistance decreases by one per cent. This may be called the power sensitivity. It is equal to $C/(\alpha \times 100)$ or about one to ten milliwatts per square centimeter in still air. Both C and the power sensitivity increase with air velocity. The dependence of C on gas pressure and velocity is the basis of the use of thermistors as manometers, and as anemometers or flowmeters. Note that in Figure 10 one curve can be superposed on any other by a shift along a constant resistance line.

Figure 11 shows a family of log V versus log I curves for various values of R_o while B, C, and T_o are kept constant. This can be brought about by changing the length, width, and thickness to vary R_o while the surface area is kept constant. If the resistance had been changed by changing the ambient temperature, T_o, the resulting curves would not appear very different from those shown. Note that one curve can be superposed on any other curve by a shift along a constant power line.

Figure 12 shows a family of log V versus log I curves for eight different values of B while C, R_o, and T_o are kept constant. In contrast to the curves in Figures 10 and 11 in which any curve could be obtained from any other curve by a shift along an appropriate axis, the curves in Figure 12 are each distinct. For each curve there exists a limiting ohmic resistance for low currents and another for high currents. For $B = 0$ these two are identical. As B becomes larger, the log of the ratio of the two limiting resistances increases proportionally to B. Note also that for $B > 1{,}200$ degrees Kelvin, the curves have a maximum. For large B values this maximum occurs at low powers and hence at low values of $T - T_o$. This follows as $W = C(T - T_o)$. As B decreases, V_m occurs at increasingly higher powers or temperatures. For $B < 1{,}200$ degrees Kelvin no maximum exists.

The curves in Figures 10 to 12 have been drawn for the ideal case in which the resistance in series with the thermistor is zero, and in which no temperature limitations have been considered. In any actual case there always is some unavoidable small resistance, such as that of the leads, in series with the thermistor, and hence the parts of the curves corresponding to low resistances may not be observable. Also, at high powers the temperature may attain such values that something in the thermistor structure will go to pieces, thus limiting the range of observation. These unobservable ranges have been indicated by dashed lines in Figure 12. The exact location of the dashed portions will of course depend on how a completed thermistor is constructed. In setting these limits we also have considered temperature limitations beyond which aging effects might become too great.

The curves in Figures 9 to 12 have been computed on the basis of the following equations:

$$R = R_o \epsilon^{(B/T)-(B/T_o)} = V/I \tag{10}$$

$$W = C(T - T_o) = VI \tag{11}$$

For these curves the constants R_o, T_o, B, and C are specified. The values of temperature, T_m, power, W_m, resistance, R_m, voltage, V_m, and current, I_m, that prevail at the maximum in the voltage current curve are given by the following equations in which T_m is chosen as the independent parameter. By differentiating equations 10 and 11 with respect to I, putting the derivatives equal to zero, one obtains

$$T_m^2 = B(T_m - T_o) \tag{12}$$

whose solution is

$$T_m = (B/2)(1 \mp \sqrt{1 - 4T_o/B}) \tag{13}$$

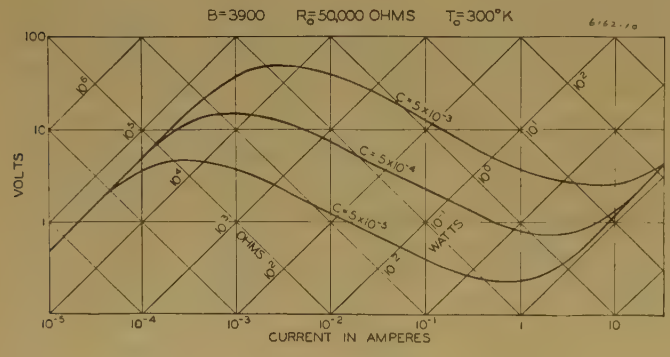

Figure 10 (above). Logarithmic plots of voltage versus current for three values of the dissipation constant C

These curves are calculated for the constants given in the upper part of the figure

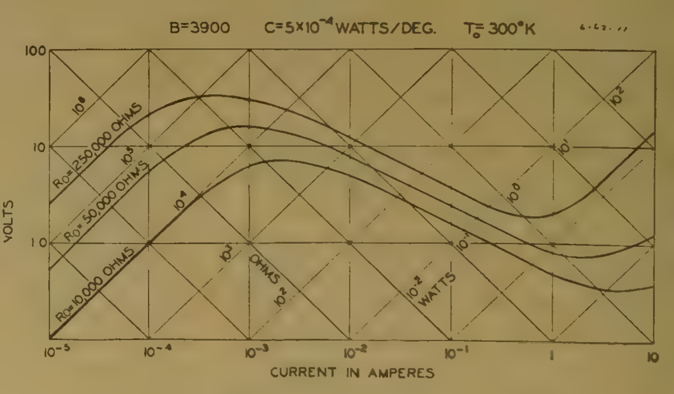

Figure 11 (right, above). Logarithmic plots of voltage versus current for three values of the resistance, R_o, at ambient temperature

These curves are calculated for the constants given in the upper part of the figure

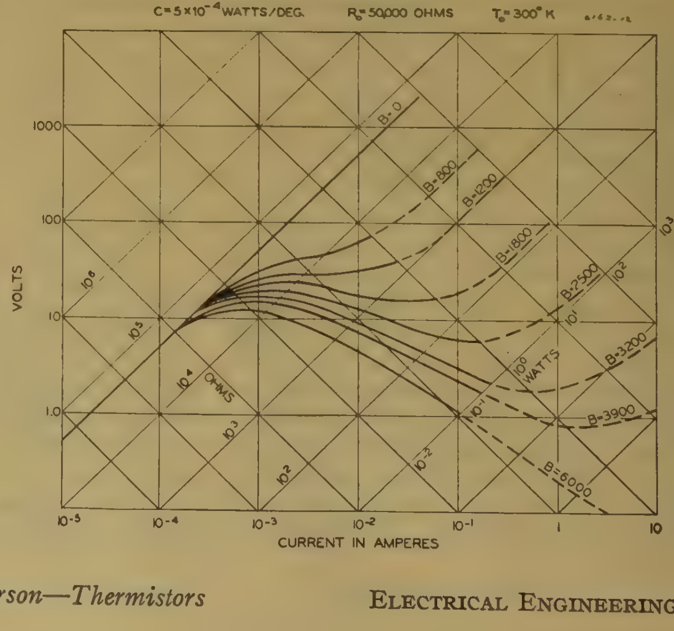

Figure 12 (right). Logarithmic plots of voltage versus current for eight values of B in equation 1

These curves are calculated for the constants given in the upper part of the figure

The minus sign pertains to the maximum in Figures 10 to 12 while the plus sign pertains to the minimum. Note that T_m depends only on B and T_o, and not on R, R_o, or C. From equations 4, 10, and 11 it follows that

$$-\alpha_m(T_m-T_o)=1 \tag{14}$$

$$W_m=C(T_m-T_o) \tag{15}$$

$$R_m=R_o\epsilon^{-T_m/T_0}\doteq R_0\epsilon^{-1}[1-(T_m-T_o)/T_o+(^1/_2)\{(T_m-T_o/T_o)\}^2-\ldots] \tag{16}$$

$$V_m=[CR_o(T_m-T_o)(\epsilon^{-T_m/T_o})]^{1/2} =\left\{CR_o(T_m-T_o)\epsilon^{-1}[1-(T_m-T_o)/T_o+(^1/_2)\{(T_m-T_o)/T_o\}^2-\ldots]\right\}^{1/2} \tag{17}$$

$$I_m=[(C/R_o)(T_m-T_o)\epsilon^{T_m/T_o}]^{1/2} =\left\{(C/R_o)(T_m-T_o)\epsilon[1+(T_m-T_o)/T_o+(^1/_2)\{(T_m-T_o)/T_o\}^2+\ \ldots]\right\}^{1/2} \tag{18}$$

Thus far the presentation has been limited to steady state conditions, in which the power supplied to the thermistor is equal to the power dissipated by it, and the temperature remains constant. In many cases, however, it is important to consider transient conditions when the temperature, and any quantities which are functions of temperature, vary with time. A simple case which will illustrate the concepts and constants involved in such problems is as follows: A massive thermistor is heated to about 150 to 200 degrees centigrade by operating it well beyond its peak. At time $t = 0$, the circuit is switched over to a constant current having a value so small that I^2R always is negligibly small. The voltage across the thermistor then is followed as a function of time. From this, the resistance and temperature are computed. Figure 13 shows a plot of log $(T - T_a)$ versus t for a rod thermistor of "material 1", about 1.2 centimeters long, 0.30 centimeter in diameter, and weighing 0.380 gram. In any time interval Δt, there are $C(T - T_a)\Delta t$ joules being dissipated. As a result the temperature will decrease by ΔT given by

$$-H\Delta T=C(T-T_a)\Delta t$$

or

$$(T-T_a)=-(H/C)(\Delta T/\Delta t) \tag{19}$$

where H = heat capacity in joules per degree centigrade. The solution of this equation is

$$(T-T_a)=(T_o-T_a)\epsilon^{-t/\tau} \tag{20}$$

in which $T_o = T$ when $t = 0$, and

$$\tau=H/C \tag{21}$$

where τ is in seconds. It commonly is called the time constant. From equation 20 it follows that a plot of log $(T - T_a)$ versus t should yield a straight line whose slope $= -1/2.303\ \tau$. If H and C vary slightly with temperature, then τ will vary slightly with T and t. The line will not be perfectly straight, but its slope at any t or $(T - T_a)$ will yield the appropriate τ or H/C for that T. As previously described, C can be determined from a plot of watts dissipated versus T. For this thermistor this curve became steeper at the higher temperatures so that C increased for higher temperatures. Table I summarizes the values of C, τ, and H at various T.

When a thermistor is heated by passing current through it, conditions are involved somewhat more since the I^2R power will be a function of time. At any time in the heating cycle, the heat power liberated will be equal to the watts dissipated, or $C(T - T_a)$ plus watts required to raise the temperature, or HdT/dt. The heat power liberated will depend on the circuit conditions. In a circuit like that shown in the upper corner of Figure 14, the current varies with time, as shown by the six curves for six values of the battery voltage E. If a relay in the circuit operates when the current reaches a definite value, a considerable range of time delays can be achieved. This family of curves will be modified by changes in ambient temperature and where rather precise time delays are required, the ambient temperature must be controlled or compensated.

Since thermistors cover a wide range in size, shape, and heat conductivity of surrounding media, large variations in H, C, and τ can be produced. The time constant can be varied from about one millisecond to about ten minutes or a millionfold.

One very important property of a thermistor is its aging characteristic, or how constant the resistance at a given temperature stays with use. To obtain a stable thermistor it is necessary to

1. Select only semiconductors which are pure electronic conductors.
2. Select those which do not change chemically when exposed to the atmosphere at elevated temperatures.
3. Select one which is not sensitive to impurities likely to be encountered in manufacture or in use.
4. Treat it so that the degree of dispersion of the critical impurities is in equilibrium, or else that the approach to equilibrium is very slow at operating temperatures.
5. Make a contact which is intimate, sticks tenaciously, has an expansion coefficient compatible with the semiconductor, and is durable in the atmospheres to which it will be exposed.
6. In some cases, enclose the thermistor in a thin coat of glass or material impervious to gases and liquids, the coat having a suitable expansion coefficient.
7. Preage the unit for several days or weeks at a temperature somewhat higher than that to which it will be subjected.

By taking these precautions, remarkably good stabilities can be obtained.

Figure 15 shows aging data taken on three-quarter inch diameter disks of "material 1" and "material 2" with silver contacts and soldered leads. These disks were measured soon after production, were aged in an oven at 105 degrees centigrade, and were tested periodically at 24 degrees centigrade. The percentage change in resistance over its initial value is plotted versus the logarithm of the time in the aging oven. It is to be noted that most of the aging takes place in the first day or week. If these disks were preaged for a week or a month and the subsequent change in resistance referred to the resistance after

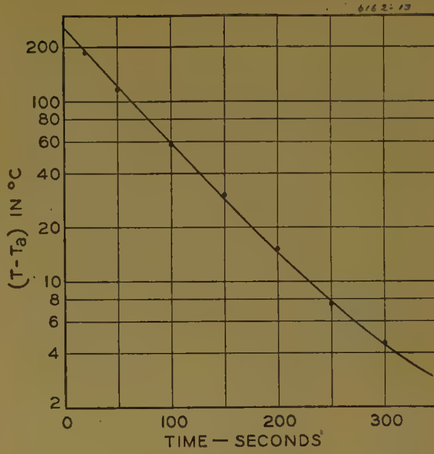

Figure 13. Cooling characteristic of a massive thermistor—logarithm of temperature above ambient versus time

Table I. Values of C, τ, H as Functions of T for a Thermistor of "Material 1"

Thermistor About 1.2 Centimeters Long' 0.30 Centimeter in Diameter, and Weighing 0.380 Gram. T_a=24 Degrees Centigrade

T, Deg C	C, Watts Per Deg C	τ, Sec	H, Joules Per Deg C	h, Joules Per Gram Per Deg C
44	0.0037	76	0.28	0.75
64	0.0037	74	0.27	0.72
84	0.0038	71	0.27	0.71
104	0.0037	69	0.26	0.68
124	0.0038	68	0.26	0.67
144	0.0038	67	0.26	0.67
164	0.0039	67	0.26	0.69
184	0.0041	66	0.27	0.71
204	0.0042	66	0.28	0.73

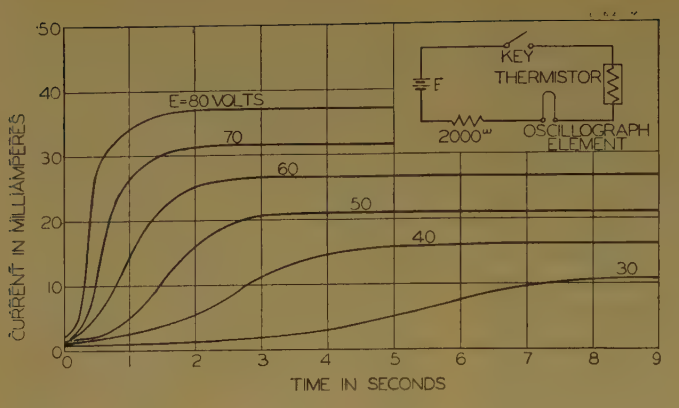

Figure 14. Current versus time curves for six values of the battery voltage in the circuit shown in the insert

preaging, they would age only about 0.2 per cent in one year. In a thermistor thermometer, this change in resistance would correspond to a temperature change of 0.05 degree centigrade. Thermistors mounted in an evacuated tube or coated with a thin layer of glass, age even less than those shown in the figure. For some applications such high stability is not essential, and it is not necessary to give the thermistors special treatment.

Thermistors have been used at higher temperatures with satisfactory aging characteristics. Extruded rods of "material 1" have been tested for stability by treating them for two months at a temperature of 300 degrees centigrade. Typical units aged from 0.5 to 1.5 per cent of their initial resistance. Similar thermistors have been exposed alternately to temperatures of 300 degrees centigrade and −75 degrees centigrade for a total of 700 temperature cycles, each lasting one-half hour. The resistance of typical units changed by less than one per cent.

In some applications of thermistors, very small changes in temperature produce small changes in potential across the thermistor, which are amplified then in high gain amplifiers. If at the same time the resistance is fluctuating randomly by as little as one part in a million, the potential across the thermistor also will fluctuate by a magnitude which will be directly proportional to the current. This fluctuating potential is called noise, and as it depends on the current, it is called current noise. In order to obtain the best signal to noise ratio, it is necessary that the current noise at operating conditions be less than Johnson or thermal noise.[7,8] To make noise-free units it is necessary to pay particular attention to the raw materials, the degree of sintering, the grain size, the method of making contact, and any steps in the process which might result in minute surface cracks or fissures.

All the thermistors discussed thus far either were heated directly by the current passing through them or by changes in ambient temperature. In indirectly heated thermistors, the temperature and resistance of the thermistor are controlled primarily by the power fed into a heater thermally coupled to it. One such form might consist of a 0.038-centimeter diameter bead of "material 2" embedded in a small cylinder of glass about 0.38 centimeter long and 0.076 centimeter in diameter. A small nichrome heater coil having a resistance of 100 ohms is wound on the glass and is fused onto it with more glass. Figure 16 shows a plot of log V versus log I for the bead element at various currents through the heater. In this way the bead resistance can be changed from 3,000 ohms to about 10 ohms. Indirectly heated thermistors ordinarily are used where the controlled circuit must be isolated electrically from the actuating circuit, and where the power from the latter must be fed into a constant resistance heater.

II—Uses of Thermistors

The thermistor, or thermally sensitive resistor, probably has excited more interest as a major electric circuit element than any other except the vacuum tube in the last decade. Its extreme versatility, small size, and ruggedness were responsible for its introduction in great numbers into communications circuits within five years after its first application in this field. The next five-year period spanned the war, and saw thermistors widely used in additional important applications. The more important of these uses ranged from time delays and temperature controls to feed-back amplifier automatic gain controls, speech volume limiters, and superhigh frequency power meters. It is surprising that such versatility can result from a temperature-dependent resistance characteristic alone. However, this effect produces a very useful nonlinear volt-ampere relationship. This, together with the ability to produce the sensitive element in a wide variety of shapes and sizes results in applications in diverse fields. The variables of design are many and interrelated, including electrical, thermal, and mechanical dimensions.

The more important uses of thermistors as indication, control, and circuit elements will be discussed, grouping the uses as they fall under the primary characteristics: resistance-temperature, volt-ampere, and current-time or dynamic relations.

Resistance–Temperature Relations

It has been pointed out in part I that the temperature coefficient of electrical resistance of thermistors is negative and several times that of the ordinary metals at room temperature. In thermistor "material 1," which commonly is used, the coefficient at 25 degrees centigrade is −4.4 per cent per degree centigrade, or over ten times that of copper, which is +0.39 per cent per degree centigrade at the same temperature. A circuit element made of this thermistor material has a resistance at zero degrees centigrade which is nine times the resistance of the same element at 50 degrees centigrade. For comparison, the resistance of a copper wire at 50 degrees centigrade is 1.21 times its value at zero degrees centigrade.

The resistance-temperature characteristics of thermistors suggest their use as sensitive thermometers, as temperature

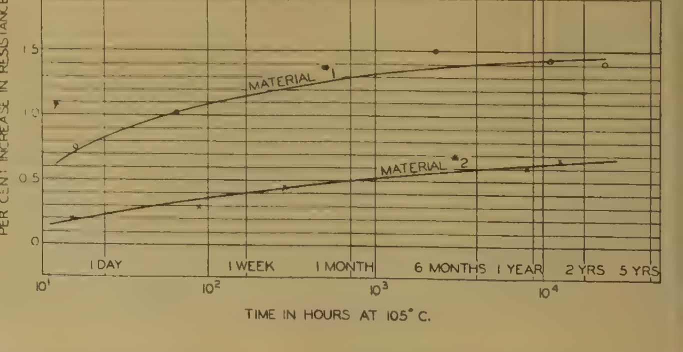

Figure 15. Aging characteristic of thermistors made of materials 1 and 2 aged in an oven at 105 degrees centigrade

Per cent increase in resistance over its initial value versus time on a logarithmic scale

actuated controls, and as compensators for the effects of varying ambient temperature on other elements in electric circuits.

Thermometry

The application of thermistors to temperature measurement follows the usual principles of resistance thermometry. However, the large value of temperature coefficient of thermistors permits a new order of sensitivity to be obtained. This and the small size, simplicity, and ruggedness of thermistors adapt them to a wide variety of temperature measuring applications. When designed for this service, thermistor thermometers have long-time stability which is good for temperatures up to 300 degrees centigrade and excellent for more moderate temperatures. A well-aged thermistor used in precision temperature measurements was found to be within 0.01 degree centigrade of its calibration after two months use at various temperatures up to 100 degrees centigrade. As development proceeds, the stability of thermistor thermometers may be expected to approach that of precision platinum thermometers. Conventional bridge or other resistance measuring circuits commonly are employed with thermistors. As with any resistance thermometer, consideration must be given to keeping the measuring current sufficiently small so that it produces no appreciable heating in order that the thermistor resistance shall be dependent upon the ambient temperature alone.

Since thermistors are readily designed for higher resistance values than metallic resistance thermometers or thermocouples, lead resistances ordinarily are not bothersome. Hence the temperature sensitive element can be located remotely

Table II. Temperature–Resistance Characteristic of a Typical Thermistor Thermometer

Temp, Deg C	Resistance, Ohms	Temp Coefficients: B, Deg C	Temp Coefficients: α, Per Cent Per Deg C
−25	580,000	3,780	−6.1
0	145,000	3,850	−5.2
25	46,000	3,920	−4.4
50	16,400	3,980	−3.8
75	6,700	4,050	−3.3
100	3,200	4,120	−3.0
150	830	4,260	−2.4
200	305	4,410	−2.0
275	100	4,600	−1.5

Dissipation constant in still air, approximately 4 milliwatts per degree centigrade; thermal time constant in still air, approximately 70 seconds; dimensions of thermistor, diameter approximately 0.11 inch, length approximately 0.54 inch.

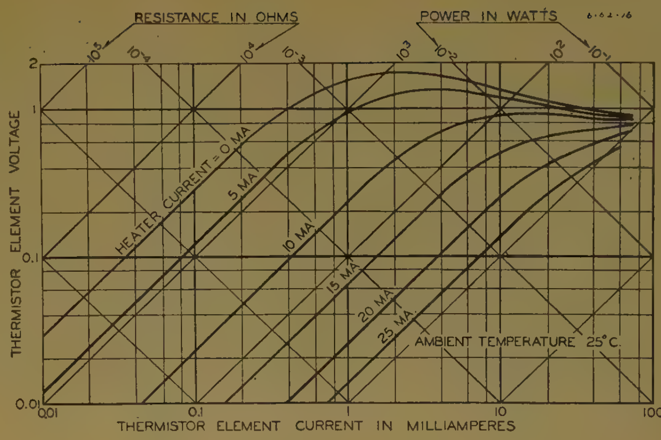

Figure 16. Logarithmic plots of voltage versus current for six values of heater current in an indirectly heated thermistor

Resistance and power scales are given on the diagonal lines

from its associated measuring circuit. This permits great flexibility in application, such, for instance, as wire line transmission of temperature indications to control points.

Table II gives the characteristics of a typical thermistor thermometer. The dissipation constant is the ratio of the power input in watts dissipated in the thermistor, to the resultant temperature rise in degrees centigrade. The time constant is the time required for the temperature of the thermistor to change 63 per cent of the difference between its initial value and that of the surroundings. As a sensitive thermometer, this thermistor with a simple Wheatstone bridge and a galvanometer whose sensitivity is 2×10^{-10} amperes per millimeter per meter will indicate readily a temperature change of 0.0005 degree centigrade. For comparison, a precision platinum resistance thermometer and the required special bridge such as the Mueller, will indicate a minimum change of 0.003 degree centigrade with a similar galvanometer.

Several thermistors which have been used for thermometry are shown in Figure 17. Included in the group are types which are suited to such diverse applications as intravenous blood thermometry and supercharger rotor temperature measurement. In Figure 17, *A* is a tiny bead with a response time of less than a second in air. *B* is a probe type unit for use in air streams or liquids. *C* is a meteorological thermometer used in automatic radio transmission of weather data from free balloons. *D* is a rod shaped unit. *E* is a disk or pellet, adapted for use in a metal thermometer bulb. Disks like the one shown have been sweated to metal plates to give a low thermal impedance connection to the object whose temperature is to be determined. *F* is a large disk with an enveloping paint finish for use in humid surroundings. The characteristics of these types are given in Table III.

The temperature of objects which are inaccessible, in motion, or too hot for contact thermometry, can be determined by permitting radiation from the object to be focused on a suitable thermistor by means of an elliptical mirror. Such a thermistor may take the form of a thin flake attached to a solid support. Its advantages compared with the thermopile and resistance bolometer are its more favorable resistance value, its ruggedness, and its high temperature coefficient of resistance. It can be made small to reduce its heat capacity so as to follow rapidly changing temperatures. Flake thermistors have been made with time constants from one millisecond to one second. Since the amount of radiant power falling on the thermistor may be quite small, sensitive meters or vacuum tube amplifiers are required to measure the small changes in the flake resistance. Where rapidly varying temperatures are not involved, thermistors with longer time constants and simpler circuit equipments can be utilized.

Temperature Control

The use of thermistors for temperature control purposes is related closely to their application as temperature measuring devices. In the ideal temperature-sensitive control element, sensitivity to temperature change should be high and the resistance value at the control temperature should be the proper value for the control circuit used. Also, the temperature rise of the control element resulting from circuit heating should be low, and the stability of calibration should be good. The size and shape of the sensitive element are dictated by several factors such as the space available, the required speed of response to temperature changes, and the amount of power which must be dissipated in the element by the control circuit to permit the arrangement to operate relays, motors, or valves.

Because of their high temperature sensitivity, thermistors have shown much

promise as control elements. Their adaptability and their stability at relatively high temperatures led, for instance, to an aircraft engine control system using a rod-shaped thermistor as the control element.[9] The thermistor, mounted in a standard one-quarter inch diameter temperature bulb assembly, operated at approximately 275 degrees centigrade. It was associated with a differential relay and control motor on the aircraft 28-volt d-c system. The power dissipation in the thermistor was two watts. The resistance of a typical thermistor under these high temperature conditions remained within ±1.5 per cent over a period of months. This corresponds to about ± one degree centigrade variation in calibration. Several other related designs were developed using the same control system with other thermistors designed for both higher and lower temperature operation. In the lower temperature applications, typical thermistors maintained their calibrations within a few tenths of a degree centigrade.

In general, gas tube control circuits dissipate less power in the thermistor than relay circuits do. This results in less temperature rise in the thermistor, and leads to a more accurate control. While the average value of this temperature rise can be allowed for in the design, the variations in different installations require individual calibration to correct the errors if they are large. The corrections may be different as a result of variations of the thermal conductivity of the surrounding media from time to time, or from one installation to another. The greater the power dissipated in the thermistor, the greater the absolute error in the control temperature for a given change in thermal conductivity. This follows from the relation

$$\Delta T = W/C \qquad (22)$$

where ΔT is the temperature rise, W is the power dissipated, and C is the dissipation constant which depends on thermal coupling to the surroundings. For the same reason, the temperature indicated by a resistance thermometer immersed in an agitated medium will depend on the rate of flow if the temperature sensitive element is operated several degrees hotter than its surroundings.

The design of a thermistor for a ventilating duct thermostat might proceed as follows as far as temperature rise is concerned:

1. Determine the power dissipation. This depends upon the circuit selected and the required over-all sensitivity.

2. Estimate the permissible temperature rise of the thermistor, set by the expected variation in air speed and the required temperature control accuracy.

3. Solve equation 22 for the dissipation constant and select a thermistor of appropriate design and size for this constant in the nominal air speed. Where more than one style of thermistor is available, the required time constant will determine the choice.

Table III. Thermistor Thermometers

	A	B	C	D	E	F
Nominal resistance, ohms:						
−25 C	—	—	87,500	610,000	—	13,000
0 C	5,000	325,000	37,500	153,000	490	3,200
25 C	2,000	100,000	18,000	48,500	175	950
50 C	900	33,000	9,700	17,300	71	340
75 C	460	13,000	5,500	7,100	32	145
100 C	250	6,000	3,700	3,400	16	70
150 C	95	1,600	—	870	4.5	—
200 C	—	500	—	—	1.6	—
300 C	—	80	—	—	—	—
Temp coefficient α, per cent per deg C at 25 C	−3.4	−4.4	−2.8	−4.4	−3.8	−4.4
Max permissible temp, deg C	150	300	100	150	200	100
Dissipation constant, C, milliwatts per deg C:						
Still air	0.1	1	7	7	—	20
Still water	—	7	—	—	—	—
Thermal time constant, sec:						
Still air	1	30	25	60	—	—
Still water	—	4	—	—	—	—
Shape	Bead	Probe	Rod	Rod	Disk	Disk
Dimensions, inches:						
Diameter or width	0.015	0.1	0.05	0.15	0.2	0.56
Length or thickness (less leads)	0.02	0.6	1.2	0.7	0.1	0.31

Compensators

It is a natural and obvious application of thermistors to use them to compensate for changes in resistance of electrical circuits caused by ambient temperature variations. A simple example is the compensation of a copper wire line, the resistance of which increases approximately 0.4 per cent per degree centigrade. A thermistor having approximately one-tenth the resistance of the copper, with a temperature coefficient of −4 per cent per degree centigrade placed in series with the line and subjected to the same ambient temperature, would serve to compensate it over a narrow temperature range. In practice however, the compensating thermistor is associated with parallel and sometimes series resistance, so that the combination gives a change in resistance closely equal and opposite to that of the circuit to be compensated over a wide range of temperatures. See Figure 18.

A copper winding having a resistance of 1,000 ohms at 25 degrees centigrade can be compensated by means of a thermistor of 566 ohms at 25 degrees centigrade in parallel with an ohmic resistance of 445 ohms as shown in Figure 18. The winding with compensator has a resistance of 1,250 ohms constant to ±1.6 per cent over the temperature range −25 degrees centigrade to +75 degrees centigrade. Over this range the copper alone varies from 807.5 ohms to 1,192.5 ohms, or ±19 per cent. The total resistance of the circuit has been increased only 6.1 per cent at the upper temperature limit by the addition of a compensator. This increase is small because of the high temperature coefficient of the compensating thermistor. The characteristics of such a thermistor are so stable that the resistance would remain constant within less than one per cent for ten years, if maintained at any temperature up to about 100 degrees centigrade. Figure 15 shows aging characteristics for typical thermistors suitable for use in compensators. These curves include the change which occurs during the seasoning period of several days at the factory, so that the aging in use is a fraction of the total shown.

In many circuits which need to function to close tolerances under wide ambient temperature variation, the values of one or more circuit elements may vary undesirably with temperature. Frequently the resultant over-all variation with temperature can be reduced by the insertion of a simple thermistor placed at an appropriate point in the circuit. This is true particularly if the circuit contains vacuum tube amplifiers. In this manner, frequency and gain shifts in communications circuits have been cancelled and temperature errors prevented in the operation of devices such, for instance, as electric meters. The change in inductance of a coil resulting from the variation of magnetic characteristics of the core ma-

terial with temperature has been prevented by partially saturating the coil with direct current, the magnitude of which is directly controlled by the resistance of a thermistor imbedded in the core. In this way the amount of d-c magnetic flux is adjusted by the thermistor so that the inductance of the coil is independent of temperature.

In designing a compensator, care must be taken to ensure exposure of the thermistor to the temperature affecting the element to be compensated. Power dissipation in the thermistor must be considered and either limited to a value which will not produce a significant rise in temperature above ambient temperature, or offset in the design.

Volt-Ampere Characteristics

The nonlinear shape of the static characteristic relating voltage, current, resistance, and power for a typical thermistor was illustrated by Figure 9. The part of the curve to the right of the voltage maximum has a negative slope, applicable in a large number of ways in electric circuits. The particular characteristic shown begins with a resistance of approximately 50,000 ohms at low power. Additional power dissipation raises the temperature of the thermistor element and decreases its resistance. At the voltage maximum, the resistance is reduced to about one-third its cold value, or 17,000 ohms, and the dissipation is 13 milliwatts. The resistance becomes approximately 300 ohms when the dissipation is 100 milliwatts. Such resistance-power characteristics have resulted in the use of thermistors as sensitive power measuring devices, and as automatically variable resistances for such applications as output amplitude controls for oscillators and amplifiers. Their nonlinear characteristics also fit thermistors for use as voltage regulators, volume controls, expandors, contactless switches, and remote control devices. To permit their use in these applications for d-c as well as a-c circuits, nonpolarizing semiconductors alone are employed in thermistors with the exception of two early types.

Power Meter

Thermistors have been used very extensively in the ultra- and superhigh frequency ranges in test sets as power measuring elements. The particular advantages of thermistors for this use are that they can be made small in size, have a small electrical capacity, can be severely overloaded without change in calibration, and easily can be calibrated with d-c or low frequency power. For this application the thermistor is used as a power absorbing terminating resistance in the transmission line, which may be of Lecher, coaxial, or wave-guide form. Methods of mounting have been worked out which reduce the reflection of high frequency energy from the termination to negligible values, and assure accurate measurement of the power over broad bands in the frequency spectrum. Conventionally, the thermistor is operated as one arm of a Wheatstone bridge, and is biased with low frequency or d-c energy to a selected operating resistance value, for instance, 125 or 250 ohms in the absence of the power to be measured. The application of the power to be measured further decreases the thermistor resistance, the bridge becomes unbalanced, and a deflection is obtained on the bridge meter. A full scale power indication of one milliwatt is customary for the test set described, although values from 0.1 milliwatt to 200 milliwatts have been employed using thermistors with different sized beads as shown in Figure 19.

Continuous operation tests of these thermistors indicate very satisfactory stability with an indefinitely long life. A group of eight power meter thermistors, normally operated at 10 milliwatts and having a maximum rating of 20 milliwatts, were operated for over 3,000 hours at a power input of 30 milliwatts. During this time the room temperature resistance remained within 1.5 per cent of its initial value, and the power sensitivity, which is the significant characteristic, changed by less than 0.5 per cent.

When power measuring test sets are intended for use with wide ambient temperature variations, it is necessary to temperature compensate the thermistor. This is accomplished conventionally by the introduction of two other thermistors into the bridge circuit. These units are designed to be insensitive to bridge currents but responsive to ambient temperature. One of the compensators maintains the zero point and the other holds the meter scale calibration independent of the effect of temperature change on the measuring thermistor characteristics.

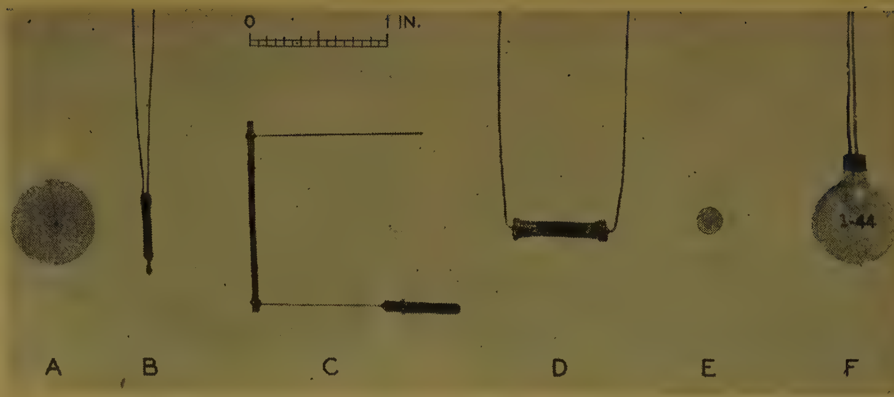

Figure 17. Some forms of thermistors which have been used as resistance thermometers

Automatic Oscillator Amplitude Control

Meacham,[10] and Shepherd and Wise[11] have described the use of thermistors to provide an effective method of amplitude stabilization of both low and high frequency oscillators. These circuits oscillate because of positive feedback around the vacuum tube. The feed-back circuit is a bridge with at least one arm containing a thermistor which is heated by the oscillator output. Through this arrangement, the feedback depends in phase and magnitude upon the output, and there is one value of thermistor resistance which, if attained, would balance the bridge and cause the oscillation amplitude to vanish. Obviously this condition never can be exactly attained, and the operating point is just enough different to keep the bridge slightly unbalanced and produce a predetermined steady value of oscillation output. Such oscillators in which the amplitude is determined by thermistor nonlinearity have manifold advantages over those whose amplitude is limited by vacuum tube nonlinearity. The harmonic content in the output is smaller, and the performance is much less dependent upon the individual vacuum tube and upon variations of the supply voltages. It is necessary that the thermal inertia of the thermistor be sufficient to prevent it from varying in resistance at the oscillation frequency. This is satisfied easily for all frequencies down to a small fraction of a cycle per second. Figure 20 shows a thermistor frequently used for oscillator control, together with its static

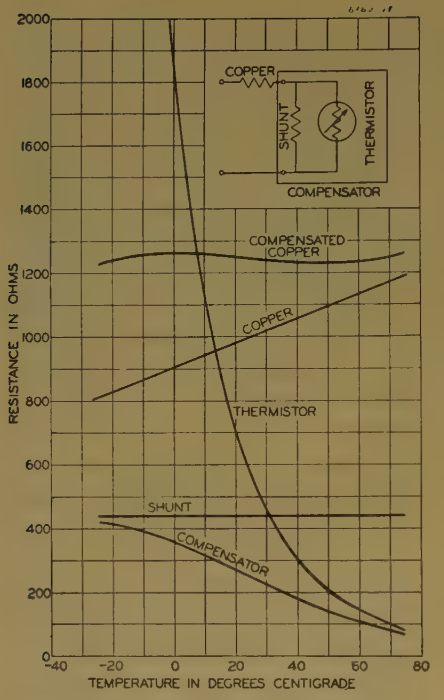

Figure 18. Temperature compensation of a copper conductor by means of a thermistor network

electrical characteristic. This thermistor is satisfactory in oscillators for frequencies above approximately 100 cycles per second. Similar types have been developed with response characteristics suited to lower frequencies, and for other resistance values and powers.

Where the ambient temperature sensitivity of the thermistor is disadvantageous in oscillator controls, the thermistor can be compensated by thermostating it with a heater and compensating thermistor network, as shown in Figure 21.

Amplifier Automatic Gain Control

Since the resistance of a thermistor of suitable design varies markedly with the power dissipated in it or in a closely associated heater, such thermistors have proved to be very valuable as automatic gain controls, especially for use with negative feed-back amplifiers. This arrangement has seen extensive use in wire communication circuits for transmission level regulation, and has been described in some detail elsewhere.[12-14] In one form, a directly heated thermistor is connected into the feed-back circuit of the amplifier in such a way that the amount of feedback voltage is varied to compensate for any change in the output signal. By this arrangement, the gain of each amplifier in the transmission system is adjusted continually to correct for variations in overall loss due to weather conditions and other factors, so that constant transmission is obtained over the channel at all times. In a carrier system now in extensive use, the system gain is regulated principally in this way. In this system the transmission loss variations due to temperature are not the same in all parts of the pass band. The loss is corrected at certain repeater points along the transmission line by two additional thermistor gain controls: slope, proportional to frequency; and bulge, with a maximum at one frequency. These thermistors are indirectly heated, with their heaters actuated by energy dependent upon the amplitude of the separate pilot carriers, which are introduced at the sending end for the purpose.

In this type of application, the thermistor will react to the ambient temperature to which it is exposed, as well as to the current passing through it. Where this is important, the reaction to ambient temperature can be eliminated by the use of a heater-type thermistor as shown in Figure 21. The heater is connected to an auxiliary circuit containing a temperature compensating thermistor. This circuit is so arranged that the power fed into the heater of the gain control thermistor is just sufficient at any ambient temperature to give a controlled and constant value of temperature in the vicinity of the gain control thermistor element.

Another interesting form of thermistor gain control utilizes a heater-type thermistor, with the heater driven by the output of the amplifier and with the thermistor element in the input circuit, as shown in Figure 22. In this arrangement, the feedback is accomplished by thermal, rather than electrical coupling. A broadband carrier system, now in use, is regulated with this type of thermistor. In this system a pilot frequency is supplied, and current of this frequency, selected by a network in the regulator, actuates the heater of the thermistor to give smooth, continuous gain control.

By utilizing a heater thermistor of different characteristics, the circuit and load of Figure 22 may be given protection against overloads. In this application, the sensitivity and element resistance of the thermistor are chosen so that the thermistor element forms a shunt of high resistance value, so as to have negligible effect on the amplifier for any normal value of output. However, if the output power rises to an abnormal level, the thermistor element becomes heated and reduced in resistance. This shunts the input to the amplifier, and thus limits the output. Choice of a thermistor having a suitable time constant permits the onset of the limiting effect to be delayed

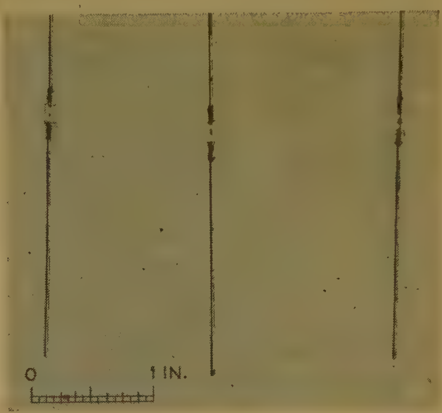

Figure 19. Power measuring thermistors with different sized beads

Figure 20A (above). An amplitude control thermistor

The glass bulb is 1.5 inches in length

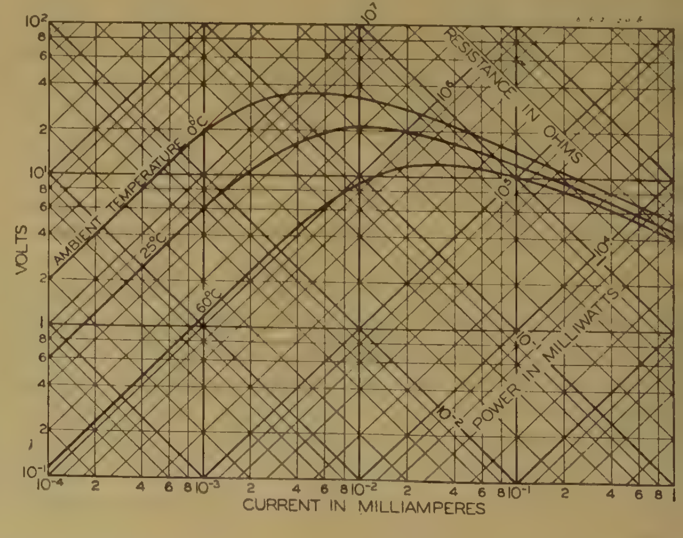

Figure 20B (right). Steady state characteristics of amplitude control thermistor shown in Figure 20A

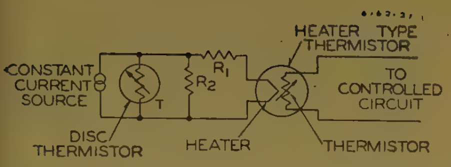

Figure 21. Circuit employing an auxiliary disk thermistor to compensate for effect of varying ambient temperature on a control thermistor

for any period from about a second to a few minutes.

Regulators and Limiters

A group of related applications for thermistors depends on their steady state nonlinear volt-ampere characteristic. These are the voltage regulator, the speech volume limiter, the compressor, and the expandor. The compressor and expandor are devices for altering the range of signal amplitudes. The compressor functions to reduce the range, while the expandor increases it. In Figure 23, curve 1 is a typical thermistor static characteristic having negative slope to the right of the voltage maximum. Curve 2 is the characteristic of an ohmic resistance R having an equal but positive slope. Curve 3 is the characteristic obtained if the thermistor and resistor are placed in series. It has an extensive segment where the voltage is almost independent of the current. This is the condition for a voltage regulator or limiter. If a larger value of resistance is used, as in curve 4, its combination with the thermistor in series results in curve 5, the compressor. In these uses the thermistor regulator is in shunt with the load resistance, so that

$$E=E_0=E_I-IR_S \tag{23}$$

in the circuit diagram of Figure 23. Here E is the voltage across the thermistor and resistor R, E_0 is the output voltage, and E_I, I and R_S are respectively, the input voltage, current, and resistance.

If the thermistor and associated resistor are placed in series between the generator and load resistance, an expandor is obtained, and

$$E_0=E_I-E \tag{24}$$

As the resistance R in series with the thermistor is increased, the degree of expansion is decreased, and vice versa.

The treatment thus far in this section assumes that change of operating point occurs slowly enough to follow along the static curves. For a sufficiently rapid change of the operating point, the latter departs from the static curve and tends to progress along an ohmic resistance line intersecting the static curve. For sufficiently rapid fluctuations, control action then may be derived from the resistance changes resulting from the rms power dissipated in the thermistor unit. In speech volume limiters, the thermistor is designed for a speed of response that will produce limiting action for the changes in volume which are syllabic in frequency or slower, and that will not follow the more rapid speech fluctuations with resulting change in wave shape or nonlinear distortion. Speech volume limiters of this type can accommodate large volume changes without producing wave form distortion.[13,16]

Remote Control Switches

The contactless switch and rheostat are natural extensions of the uses just discussed. The thermistor is used as an element in the circuit which is to be controlled, while the thermistor resistance value is, in turn, dependent upon the energy dissipated directly or indirectly in it by the controlling circuit. By taking advantage of the nonlinearity of the static volt-ampere characteristic, it is possible to provide snap and lock-in action in some applications.

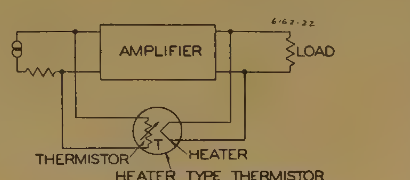

Figure 22. Thermal feed-back circuit for gain control purposes

This arrangement has also been used as a protective circuit for overloads

Manometer

Several interesting and useful applications such as vacuum gauges, gas analyzers, flowmeters, thermal conductivity meters, and liquid level gauges of high sensitivity and low operating temperature are based upon the physical principle that the dissipation constant of the thermistor depends on the thermal conductivity of the medium in which it is immersed. As shown in Figure 10, a change in this constant shifts the position of the static characteristic with respect to the axes. In these applications, the undesired response of the thermistor to the ambient temperature of the medium can, in many cases, be eliminated or reduced by introducing a second thermistor of similar characteristics into the measuring circuit. The compensating thermistor is subjected to the same ambient temperature, but is shielded from the effect being measured, such as gas pressure or flow. The two thermistors can be connected into adjacent arms of a Wheatstone bridge which is balanced when the test effect is zero, and becomes unbalanced when the effective thermal conductivity of the medium is increased. In gas flow measurements, the minimum measurable velocity is limited, as in all hot wire devices, by the convection currents produced by the heated thermistor.

The vacuum gauge or manometer which is typical of these applications will be described somewhat in detail. The sensitive element of the thermistor manometer is a small glass coated bead 0.02 inch in diameter, suspended by two fine wire leads in a tubular bulb for attachment to the chamber whose gas pressure is to be measured. The volt-ampere characteristics of a typical laboratory model manometer are shown in Figure 24 for air at several absolute pressures from 10^{-6} millimeters of mercury to atmospheric pressure. The operating point, in general, is to the right of the peak of these curves. Electrically this element is connected into a unity ratio arm Wheatstone bridge with a similar, but evacuated, thermistor in an adjacent arm as shown in the circuit schematic of Figure

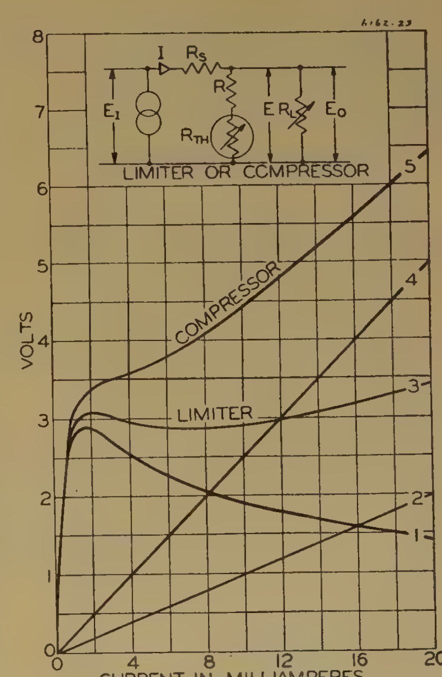

Figure 23. Characteristics of a simple thermistor voltage regulator, limiter, or compressor circuit

25. The air pressure calibration for such a manometer is also shown. The characteristic will be shifted when a gas is used having a thermal conductivity different from that of air. Such a manometer has been found to be best suited for the measurement of pressures from 10^{-5} to 10 millimeters of mercury. The lower pressure limit is set by practical considerations, such as meter sensitivity and the ability to maintain the zero setting for reasonable periods of time in the presence of the variations of supply voltage and ambient temperature. The upper pressure measurement limit is caused by the onset of saturation in the bridge unbalance voltage versus pressure characteristic at high pressures. This is basically because the mean free path of the gas molecules becomes short compared with the distance between the thermistor bead and the inner surface of the manometer bulb, so that the cooling effect becomes nearly independent of the pressure.

The thermistor manometer is specially advantageous for use in gases which may be decomposed thermally. For this type of use, the thermistor element temperature can be limited to a rise of 30 degrees centigrade or less above ambient temperature. For ordinary applications, however, a temperature rise over a range not exceeding 200 degrees centigrade in vacuum, permits measurement over wider ranges of pressure. Special models also have been made for use in corrosive gases. These expose only glass and platinum alloy to the gas under test.

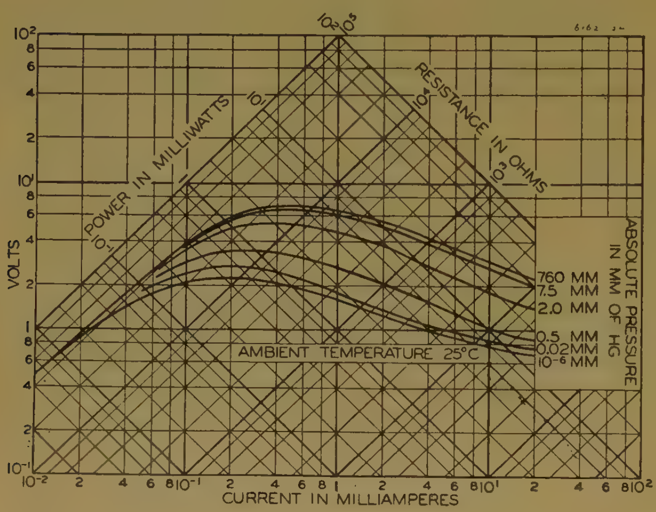

Figure 24 (left). Characteristics of a typical thermistor manometer tube, showing the effect of gas pressure on the volt-ampere and resistance-power relations

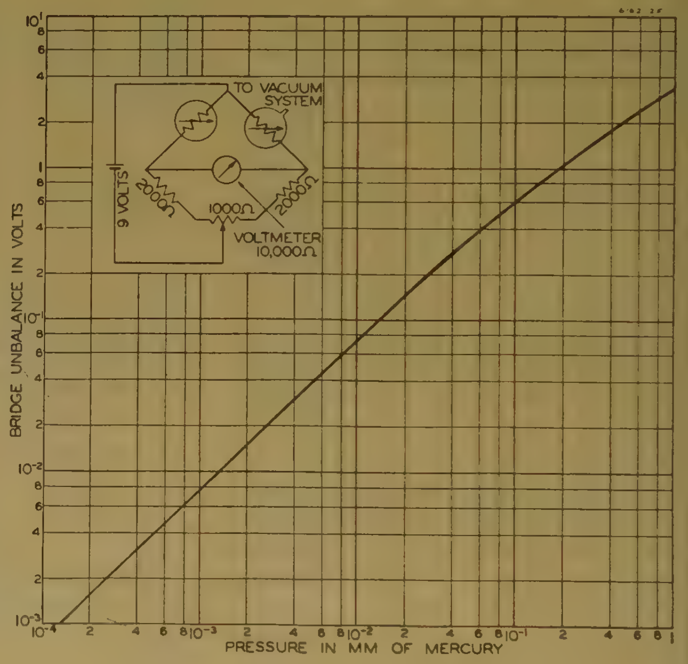

Figure 25 (right). Operating circuit and calibration for a vacuum gauge utilizing the thermistor of Figure 24

Timing Devices

The numerically greatest application for thermistors in the communication field has been for time delay purposes. The physical basis for this use has been discussed in part I for the case of a directly heated thermistor placed in series with a voltage source and a load, to delay the current rise after circuit closure. This type of operation will be termed the power driven time delay.

By the use of a thermistor suited to the circuit and operating conditions, power driven time delays can be produced from a few milliseconds to the order of a few minutes. Thermistors of this sort have the advantage of small size, light weight, ruggedness, indefinitely long life, and absence of contacts, moving parts, or pneumatic orifices which require maintenance care. Power driven time delay thermistors are best fitted for applications where close limits on the time interval are not required. In some communications uses, it is satisfactory to permit a six to one ratio between maximum and minimum times as a result of the simultaneous variation from nominal values of all the following factors which affect the delay: operating voltage ± five per cent; ambient temperature 20 degrees centigrade to 40 degrees centigrade; operating current of the relay ±25 per cent; relay resistance ± five per cent; and thermistor variations such as occur from unit to unit of the same type.

After a timing operation, a power driven time delay thermistor should be allowed time to cool before a second operation. If this is not done, the second timing interval will be shorter than the first. The cooling period depends on particular circuit conditions and details of thermistor design, but generally is several times the working time delay. In telephone relay circuits requiring a timing operation soon after previous use, the thermistor usually is connected so that it is short circuited by the relay contacts at the close of the working time delay interval. This permits the thermistor to cool during the period when the relay is locked up. If this period is sufficiently long, the thermistor is available for use as soon as the relay drops out. Time delay thermistors have been operated more than half a million times on life test, with no significant change in their timing action.

To avoid the limitations of wide timing interval limits and extended cooling periods between operations usually associated with the power driven time delay thermistor, a cooling time delay method of operation has been used. In this arrangement, two relays or the equivalent are employed, and the thermistor is heated to a low resistance value by passing a relatively large current through it for an interval, short compared with the desired time interval. The current then is reduced automatically to a lower value, and the thermistor cools until its resistance increases enough to reduce the current further and trip the working relay.

This part of the operating cycle accounts for the greater part of the desired time interval. With this arrangement, the thermistor is available for reuse immediately after a completed timing interval, or, as a matter of fact, after any part of it. By proper choice of operating currents and circuit values, wide variations of voltage and ambient temperature may occur with relatively little effect upon the time interval. The principal variable left is the cooling time of the thermistor itself. This is fixed in a given thermistor unit, but may vary from unit to unit, depending upon dissipation constant and thermal capacity, as pointed out above.

In addition to their use as definite time delay devices, thermistors have been used in several related applications. Surges can be prevented from operating relays or disturbing sensitive apparatus by introducing a thermistor in series with the circuit component which is to be protected. In case of a surge, the high initial resistance of the thermistor holds the surge current to a low value, provided that the surge does not persist long enough to overcome the thermal inertia of the thermistor. The normal operating voltage, on the other hand, is applied long enough to lower the thermistor resistance to a negligible value, so that a normal operating current will flow after a short interval. In this way, the thermistor enables the circuit to distinguish between an undesired signal of short duration and a desired signal of longer duration, even though the undesired impulse is several times higher in voltage than the signal.

Oscillators, Modulators, and Amplifiers

A group of applications, already explored in the laboratory but not put into engineering use, includes oscillators, modulators, and amplifiers for the low and audio-frequency range. If a thermistor is biased at a point on the negative slope portion of the steady-state volt-ampere characteristic, and if a small alternating voltage is then superposed on the direct voltage, a small alternating current will flow. If the thermistor has a small time constant, τ, and if the applied frequency is low enough, the alternating volt-ampere characteristic will follow the steady-state curve and dV/dI will be negative. As the frequency of the applied alternating voltage is increased, the value of the negative resistance decreases. At some critical frequency, f_c, the resistance is zero and the current is 90 degrees out of phase with the voltage. In the neighborhood of f_c, the thermistor acts like an inductance whose value is of the order of a henry. As the frequency is increased beyond f_c, the resistance is positive, and increases steadily until it approaches the d-c value when the current and voltage are in phase. The critical frequency is approximately given by

$$f_c = {}^1/_{2\tau} \qquad (25)$$

If τ can be made as small as 5×10^{-5} seconds, f_c is equal to 10,000 cycles per second, and the thermistor would have an approximately constant negative resistance up to half this frequency. Point contact thermistors, having such critical frequencies or even higher, have been made in a number of laboratories. However, none of them has been made with sufficient reproducibility and constancy to be useful to the engineer. It has been shown both theoretically and experimentally that any negative resistance device can be used as an oscillator, a modulator, or an amplifier. With further development, it seems probable that thermistors will be used in these fields.

Summary

The general principles of thermistor operation, and examples of specific uses have been given to facilitate a better understanding of thermistors, with the feeling that such an understanding will be the basis for increased use of this new circuit and control element in technology.

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Characteristics of Resistance Welding Transformers

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RESISTANCE WELDING as a labor saving production method has made tremendous strides in the past decade. Its widespread recognition during World War II has made it quite evident that this art will be used even more universally in the postwar era.

When one considers that the heart of the resistance welding machine is its transformer, it appears quite logical that an explanation of this type transformer is forthcoming.

It is the purpose of this paper to present the characteristics and functions of the resistance welding transformer in addition to the design problems encountered.

The subject is much too large to enable a detailed discussion and derivation of all the mathematics required to determine the design constants used in this field. Emphasis has been placed on the general problems of design and the effects of deviations from standard practice.

Functions of a Resistance Welding Transformer

The resistance welding transformer, in common with all transformers, is a device for transferring energy by means of electromagnetic induction.

The function of a resistance welding transformer is to furnish relatively large currents at comparably low voltages from the mains of its user's power supply. When delivering load the transformer can be considered, for all practical purposes, as under short-circuit conditions. Its current is limited only by the impedance of the welder plus a small impedance caused by the work.

Standards Governing Resistance Welding Transformers

The Resistance Welder Manufacturers Association has set up standards covering transformers used in resistance welding machines.

The capacity of resistance welding transformers is rated in kilovolt-amperes at 50 per cent duty cycle. The continuous duty rating of such transformers then becomes 0.707 times the name plate rating.

The duty cycle expressed in per cent is considered to be the ratio of weld time in cycles per minute to the total cycles in a minute times 100.

$$\text{Per cent duty cycle on 60 cycles} = \frac{\text{cycles of weld time in a minute}}{3{,}600} \times 100 \qquad (1)$$

As the users of resistance welding equipment usually are more concerned with the permissible duty cycle for a given load than they are with the name plate rating, it often is desired to convert as follows:

$$\text{Permissible duty cycle in per cent} = \left(\frac{\text{rated kilovolt-amperes} \times 7.07}{\text{load kilovolt-amperes}}\right)^2 \qquad (2)$$

These same standards require that the exciting currents be limited in proportion to the rated current. Transformers rated at less than 100 kva are permitted an exciting current of ten per cent of the rated current, while those rated 100 kva or over are limited to exciting currents of only five per cent of the rated current.

Insulations are required to be able to withstand twice the maximum induced voltage plus 1,000 volts for a period of one minute.

The permitted temperature rise is governed by the type of insulation used. Transformers in which class B insulation is used are limited to a temperature rise of 80 degrees centigrade, while those in which class A insulation is used are permitted only 60 degrees centigrade rise, as measured by the thermocouple method.

Design Restrictions

The design of resistance welding transformers can be approached only after consideration of several points of limitation.

1. The geometry of the transformer frequently is determined by the available space provided in the welding unit.

2. The limitations as set forth by the RWMA must be met.

3. It is desired to minimize the losses.

4. Economics dictate the manufacture of equipment at a reasonable cost.

5. The number of secondary turns usually is limited to one.

Core Design

The use of high-grade low-watt-loss silicon steel permits the use of higher flux densities for the same hysteresis losses.

Flux densities of from 75,000 to 85,000 lines per square inch (depending on the kilovolt-ampere rating) may be used when it is desired to limit the exciting currents to those values as set forth by the RWMA

From the general formula,

$$E = 4.44NfKBA \times 10^{-8} \qquad (3)$$

and assuming a stacking factor of 0.95,

$$B = 395{,}000\frac{E}{NA} \qquad (4)$$

when $f = 60$ cycles per second.

$$\frac{A}{\frac{E}{N}} = \frac{395{,}000}{B} \qquad (5)$$

This formula gives the area of required core in square inches per volt per turn for a given flux density. Note Figure 1 for flux density versus area.

The core dimensions can be proportioned to suit the available space within certain limitations. A core which takes the form of a square results in a mean turn of minimum length and is the most economical design.

Where space dictates the use of other proportions, the length should not exceed the height by a factor of more than three to one.

Design of the High Voltage Winding

The high voltage winding usually is arranged with eight taps for heat adjustment. Where electronic controls with incorporated heat adjustment facilities are provided, a series-parallel winding without taps is becoming more popular.

The principal advantage of the series-parallel winding is increased duty cycle when less than the maximum load is required. This is true because the copper cross section remains constant as the heat is phased back.

When resistance welding transformers are provided with taps for heat adjustment the copper cross section of the high voltage winding may be reduced as the

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square of the open circuit volts across the low voltage winding. In practical design this method of calculating copper densities should be used as a minimum with space dictating how closely it is approached.

Tapering the copper cross section in this manner keeps the permissible duty cycle constant for any tap. Mathematical analysis of this will be shown.

Duty cycle is a function of kilovolt-ampere demand and rating. Then, given a welder of known kilovolt-amperes and duty cycle, it is desired to maintain the duty cycle as the load decreases.

$$\text{Demand kilovolt-amperes} = \frac{E_s I_s}{1000} \quad (6)$$

The low voltage winding current varies directly as its open circuit volts; hence,

$$\frac{E_{s1}}{E_{s2}} = \frac{I_{s1}}{I_{s2}} \quad (7)$$

$$I_{s2} = \frac{I_{s1} E_{s2}}{E_{s1}} \quad (8)$$

$$KVA_1 = \frac{E_{s1} I_{s1}}{1000} \quad (9)$$

$$KVA_2 = \frac{E_{s2} I_{s2}}{1000} \quad (10)$$

Divide one by the other:

$$\frac{KVA_1}{KVA_2} = \frac{E_{s1} I_{s1}}{E_{s2} I_{s2}} \quad (11)$$

Substitute and simplify:

$$\frac{KVA_1}{KVA_2} = \frac{E_{s1} I_{s1} E_{s1}}{E_{s2} I_{s1} E_{s2}} = \frac{E_{s1}^2}{E_{s2}^2} \quad (12)$$

From experience gathered from data on many welding transformers, the following copper densities may be used without exceeding permissible temperature rises for the two types shown:

1. For a pancake type of winding pressed tight against a water-cooled secondary winding insulated from primary winding with mica barriers, a density of 2,500 amperes per square inch may be used at 50 per cent duty cycle rating of transformer.

2. For a layer type of winding, air-cooled and used with a secondary winding not water-cooled, the permissible copper density is 1,500 amperes per square inch. This type of transformer usually has the low voltage winding or secondary made of a laminated copper band built up to the required thickness. The impedance is higher than the pancake type of winding.

Density of Low Voltage or Secondary Winding

The low voltage winding generally is made of cast copper having a minimum of 85 per cent conductivity. Water-cooling pipes are cast into or welded to the outside of the casting. When pipes are cast integral with the secondary, the casting usually is increased in thickness to allow for pipe. When copper tubing is welded to the outside of the secondary section, the thickness can be reduced to the diameter of the tubing. The impedance, therefore, is lowered by using thin sections approximately one-half inch thick and more of them than by the use of a thick section. Under ordinary conditions of cooling water temperature, the density of current in the secondary casting can be 3,200 amperes per square inch at a 50 per cent duty cycle rating.

The secondary winding current will tend to crowd at the corners and the mean turn with respect to current flow will be slightly less than when figured in the conventional manner. The calculated resistance, therefore, will give higher copper loss than actual measurement during impedance test.

Design of 100-Kva Transformer

This transformer is a 440-volt 60-cycle type with 50 per cent duty cycle. The required open circuit volts (E_s) is 6.67 with 1-turn low voltage winding.

Core Size

$B = 77{,}000$ lines per square inch (assumed)

$$A = \frac{E}{N} \times \frac{395{,}000}{B} = \frac{6.67}{1} \times \frac{395{,}000}{77{,}000} = 34.2 \text{ square inches} \quad (13)$$

High Voltage or Primary Winding

$$I_p(\text{rated}) = \frac{100{,}000}{440} = 227 \text{ amperes} \quad (14)$$

$$\text{Copper section} = 227 \times \frac{1}{2{,}500} = 0.0908 \text{ square inch} = 90{,}800 \text{ square mils} \quad (15)$$

$$\text{Number of turns for maximum } E_s = \frac{440}{6.67} = 66 \quad (16)$$

The high voltage winding will be proportioned for a series-parallel arrangement as shown in Figure 11.

The insulation between turns for a class B design can be any one of several materials. In this case the authors would choose 0.020-inch mica.

The width of primary winding copper should be kept as small as practical with one-fourth inch as the minimum. An increase in width will increase the copper loss. A practical width for this design would be 0.625 inch. As the primary winding is split into two sections, the required copper per section is

$$\frac{90{,}800}{2} \text{ square mils} = 45{,}400 \text{ square mils} = 0.0454 \text{ square inch} \quad (17)$$

The thickness of the copper per turn is $\frac{0.0454}{0.625} = 0.0728$ inch. Use 0.075 inch × 0.625 inch for practical reasons.

Therefore, the total height of high voltage winding = (0.075 + 0.020) × 33 = 3 13 inches.

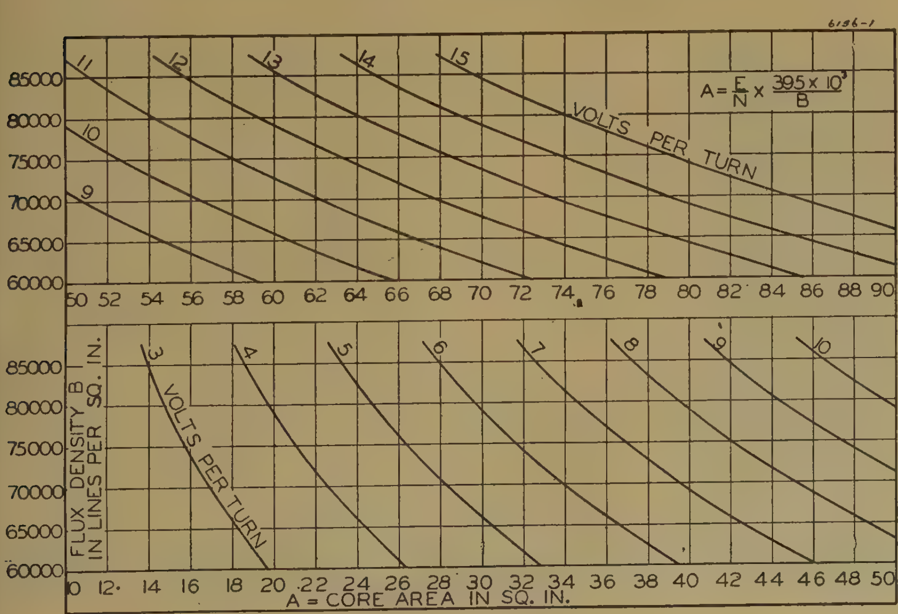
Low Voltage or Secondary Winding

$$I_s \text{ (rated)} = \frac{100{,}000}{6.67} = 15{,}000 \text{ amperes} \quad (18)$$

$$\text{Copper section} = \frac{15{,}000}{3{,}200} = 4.68 \text{ square inches} \quad (19)$$

The secondary winding will be one-half

Figure 1. Core area versus flux density



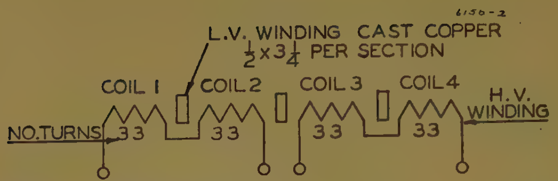


Figure 2. Schematic diagram of transformer winding

0.075×0.625 copper per turn
0.020×0.687 mica between turns

Weight = [(19 1/4×10 7/8) − (7 1/2×5)]× 5 7/8×0.241 = 242 pounds

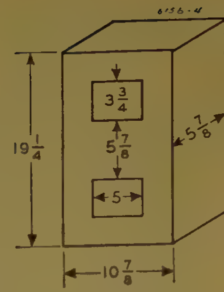


Figure 4. Detail of core

inch wide per section. The area per section = 4.68/3 = 1.56 square inch.

Height of section = 1.56 ÷ 1/2, or 3.12 inches.

Leakage Reactance*

Coil 1 (With Secondary Winding on One Side)[1]

$$L=3.2\times T_1^2\times\frac{MT}{l}\left(S+\frac{d_1+d_2}{3}\right)10^{-8} \qquad (20)$$

$$=3.2\times(33)^2\times$$

$$\frac{37}{3.25}\left(0.125+\frac{0.625+0.500}{3}\right)10^{-8}$$

$$=39{,}700\times0.500\times10^{-8}$$

$$=19{,}850\times10^{-8}\text{ henries}$$

Coil 2 (With Secondary Winding on Each Side)

$$L=3.2\times T_1^2\times\frac{MT}{l}\left(S+\frac{d_1+d_2}{6}\right)10^{-8} \qquad (21)$$

$$=39{,}700\times0.3125\times10^{-8}$$

$$=12{,}406\times10^{-8}\text{ henries}$$

Total L for coils 1 and 2

$$=(19{,}850+12{,}406)10^{-8}=32{,}256\times10^{-8}$$

$$\text{Total } L \text{ for transformer}=\frac{32{,}256\times10^{-8}}{2}$$

$$=16{,}128\times10^{-4}\text{ henries}$$

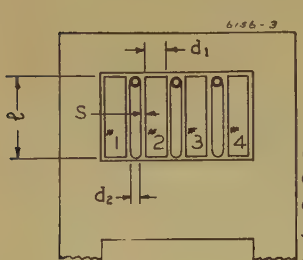
$$X_L=2\pi fL=377\times16{,}128\times10^{-8}$$

$$=0.0608\text{ ohm} \qquad (22)$$

Primary Winding Resistance

Mean turn = 37 inches = l_1
Number of turns per coil = 33 = T_1
Area of copper = (0.075×0.625)(10)[6]×1.27 = 59,500 circular mils

$$R_p=\frac{\rho_1 l_1 T_1}{A\times12}=\frac{10.37\times37\times33}{59{,}500\times12}=0.0177\text{ ohm} \qquad (23)$$



l = 3.25
d_1 = 0.625
d_2 = 0.500
S = 0.125
Mean turn = 37

Figure 3. Detail of coil arrangement

Resistance of two coils in series

$$=2\times0.0177=0.0354\text{ ohm}$$

Resistance of series-parallel winding

$$=\frac{0.0354}{2}=0.0177\text{ ohm}$$

ρ_1 = 10.37 ohms per circular-mil foot

Secondary Winding Resistance 85 Per Cent Conductivity

Mean turn = 37 inches = l_1
A = 1/2×3 1/4×3 = 4.87 square inches
= 6,190,000 circular mils

$$R_s=\frac{\rho_2 l_1}{A\times12}=\frac{10.37\times37}{0.85\times12\times6{,}190{,}000}$$

$$=6.08\times10^{-6}\text{ ohms} \qquad (24)$$

$$\rho_2=\frac{10.37}{0.85}\text{ ohms per circular-mil foot}$$

(cast copper)

Equivalent Resistance Referred To Primary Winding

$$R_{eq}=0.0177+\left(\frac{66}{1}\right)^2 6.08\times10^{-6}$$

$$=0.0177+0.0265=0.0442\text{ ohm}$$

Equivalent a-c resistance (caused by skin effect) = 1 1/2×0.0442 = 0.0663 ohm

$$Z_p=\sqrt{R_{eq}^2+X_L^2}=\sqrt{(0.0663)^2+(0.0608)^2}$$

$$=0.09\text{ ohm} \qquad (25)$$

I_pZ_p at rated current of 227A

$$=227\times0.09=20.43\text{ volts} \qquad (26)$$

$$\text{Per cent } Z_p=\frac{20.43}{440}\times100\text{ per cent}$$

$$=4.64\text{ per cent} \qquad (27)$$

$$\text{Power factor}=\frac{R_{eq}}{Z_p}=\frac{0.0663}{0.09}=0.736 \qquad (28)$$

I^2R loss at rated load = (227)[2]0.0663

$$=3{,}420\text{ watts} \qquad (29)$$

Core Loss

Core loss (note Figure 4) at 77,000 lines per square inch equals 0.9 watt per pound from curves furnished by steel manufacturers.

Total core loss = 242×0.9 = 218 watts (30)

Total loss = 3,420+218 = 3,638 watts (31)

* See Figure 3.

$$\text{Efficiency at rated load}=\frac{100{,}000}{100{,}000+3{,}638}$$

$$=0.965 \qquad (32)$$

Summary

1. Flux densities may be higher than those used in power and distribution transformers.

2. Losses caused by exciting currents have little effect on the over-all efficiency of the welding unit.

3. Losses can be minimized by the use of narrow sections.

4. Variations from optimum conditions can be made without serious effect on over-all efficiency.

5. Pancake type windings are the most efficient for resistance welding transformers.

6. Economics dictate the design of welding transformers having from three to five per cent impedance. The benefits of lower impedance transformers are offset by a greater manufacturing cost.

7. Although no mention has been made of transformers designed for frequencies other than 60 cycles, the same procedure may be followed once the proper flux density is determined.

8. The use of better insulating materials in the future will have a definite effect on present design constants.

Nomenclature

E = volts
N = turns
f = frequency
K = stacking factor
B = flux density in lines per square inch
A = area in square inches
E_s = open circuit volts of low voltage winding
I_s = current in low voltage winding
I_p = current in high voltage winding
L = inductance in henries
T_1 = turns in each high voltage coil
MT = mean turn
l = height of high voltage coil in inches
S = distance between high and low voltage windings
d_1 = width of each high voltage coil in inches
d_2 = width of low voltage coil in inches
X_L = inductive reactance in ohms
ρ_1 = resistivity of copper in ohms per circular mil foot
ρ_2 = resistivity of cast copper in ohms per circular mil foot
R_s = d-c resistance of low voltage winding
R_p = d-c resistance of high voltage winding
R_{eq} = equivalent resistance referred to the primary winding
Z_p = effective impedance referred to the primary winding

Reference

1. Electrical Machine Design (book), Alexander Gray. McGraw-Hill Book Company, New York, N. Y., 1913, page 445.

4,370,000-Kva Short-Circuit Tests on Grand Coulee 230-Kv Bus

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Synopsis: The unusually large concentration of generating capacity at Coulee Dam makes possible short-circuit power on the 230-kv bus considerably in excess of the rupturing capacity of circuit breakers now available at that voltage. Methods of sectionalizing the bus and segregating the transmission lines to limit the maximum short circuit to less than 3,500,000 kva are described in this paper. A group of fault tests on one of the 230-kv 3-cycle 3,500,000-kva steel tank oil circuit breakers was made with a maximum of six generators connected. Three-phase faults close to the breaker rating and single phase-to-ground short-circuit currents equivalent to 4,370,000-kva 3-phase were cleared with no apparent effort by the breaker. These are believed to be the heaviest short circuits ever interrupted on any power system in the world.

THE GRAND COULEE DAM and power plant is situated on the Columbia River approximately 100 miles west of the city of Spokane, in the heart of the state of Washington. At this point on the river the mean annual runoff is 79,000,000 acre-feet, the mean annual flow is 109,000 second-feet, and the dam provides a mean hydraulic head of 330 feet for operation of the turbines. The power plant is near the center of the vast transmission network of the Northwest Power Pool, which includes 11 major interconnected private and government-owned systems operating in the states of Washington, Oregon, Idaho, Montana, and Utah.

The ultimate power plant installation is to consist of two similar powerhouses, on either side of the river, each housing nine units of 108,000-kw capacity, or a combined total capacity of nearly 2,000,000 kw. The immense size of these waterwheel generators, the largest in service anywhere, is indicated by Figure 1 showing one of the rotors being lowered into place. Associated with the power plant development is an irrigation pumping plant which is situated immediately upstream from the west powerhouse and dam abutment. This pumping development ultimately will consist of 12 65,000-horsepower pumps which will lift the waters of Lake Roosevelt to an irrigation system in the Columbia Basin. Operation of these 12 pumps will be from the first six main unit generators, L-1 to L-6, through six 5,000-ampere 13.8-kv busses. Because of provisions being made for considerable water storage in the canal system, it will be possible to use these first six generating units for pumping purposes during off-peak load periods, and for commercial load at other times. Figure 2 shows a single-line diagram of the west power plant only, with dotted lines indicating the pump installations.

Such a large concentration of generating capacity introduces unique problems in circuit breaker application and requires special switching arrangements to keep the breaker duty within the rated interrupting capacity. The high speed 3-cycle 230-kv breakers on all outgoing line positions at Coulee Dam have an interrupting rating of 3,500,000 kva, and the 8-cycle generator breakers have an interrupting rating of 2,500,000 kva. The maximum interrupting rating available at the time these breakers were purchased was 2,500,000 kva. The line position breakers subsequently were rebuilt to 3,500,000 kva when interrupters of this increased rating became available.

The power plant installation at the time the fault tests were made in December 1945 consisted of six permanent 108,000-kw units and two smaller 75,000-kw units which had been borrowed temporarily from Shasta Dam, California Central Valley Project, to augment the capacity at Coulee Dam during the war. System calculations of short-circuit currents during a line fault immediately adjacent to the Coulee switchyard indicated that the 3-cycle breaker rating of 3,500,000 kva can be exceeded if more than four units and three lines are connected to the faulted section at any one time. Because of this consideration, it was deemed inadvisable to operate with more than four units on a bus section, and since an adequate synchronizing tie is necessary between the two bus sections for system stability, the interleaved scheme of operation was adopted. Figure 3 shows a simplified diagram of the interleaved line and bus arrangement showing the synchronizing ties provided by parallel line operation to the three load centers: Spokane, Midway, and Columbia. It is not the purpose of this paper to discuss in detail the interleaved scheme and its relation to system stability on the Bonneville Power Administration system, as this was presented previously.[1] However, it should be pointed out that circuit breaker capacity may dictate to a large extent the switching arrangement and method of operation on high capacity systems.

For an example of how limitations in circuit breaker capacity affect switching operations and station capacity, let us suppose that it is desired to overhaul a generator breaker, thus making it necessary to transfer the corresponding unit to

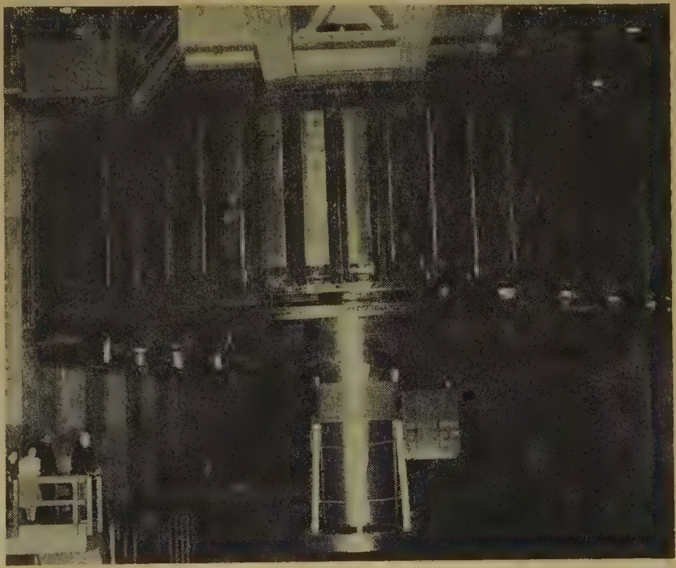

Figure 1. Rotor of 108,000-kva generator being lowered into place at Grand Coulee Dam

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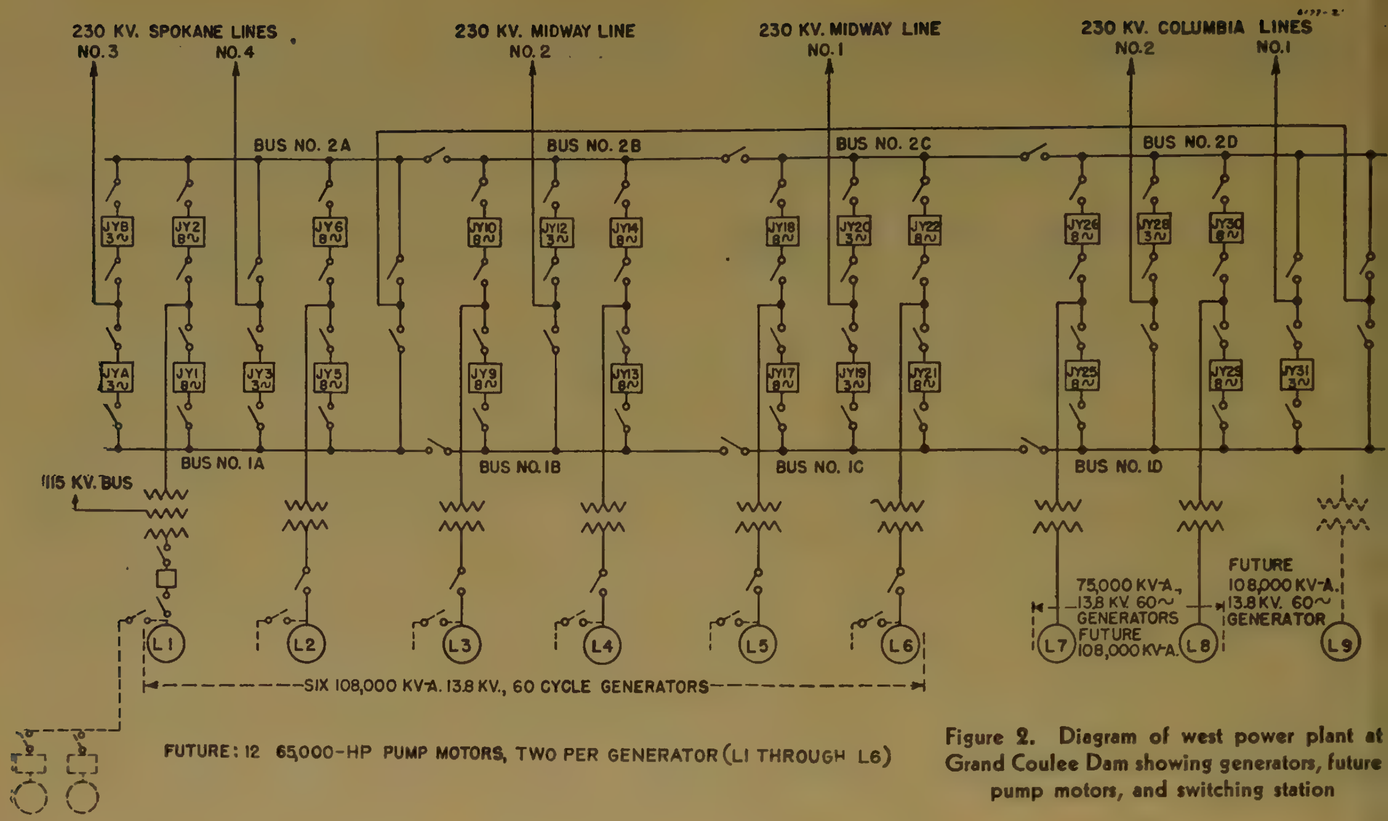

Figure 2. Diagram of west power plant at Grand Coulee Dam showing generators, future pump motors, and switching station

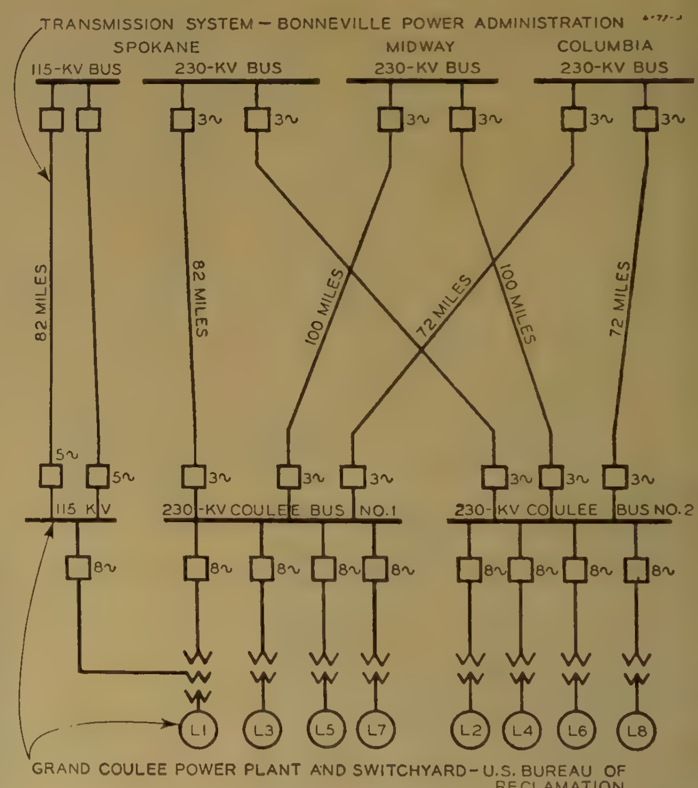

Figure 3. Simplified diagram of interleaved bus arrangement with bus ties through lines to distant stations

the opposite bus. The first operation would be to unload the unit and open the generator breaker. Before the unit can be placed on the opposite bus, however, a unit on the second bus also must be unloaded and removed from the bus as five units would exceed the interrupting rating of the breakers. This operation requires that 25 per cent of the present plant capacity, or a possible 216,000 kw of capacity must be removed from the system to accomplish the desired switching. Breakers of 5,000,000 kva capacity would make it unnecessary to remove more than one unit at a time.[2]

Fast relaying and circuit interruption during faults play an extremely important role in system stability where there are large concentrations of generating capacity and transmission lines that are heavily loaded. This is accentuated on the system under discussion by the fact that there are lumped loads and lumped generating capacity in the Pacific Northwest, separated by long transmission distances. With over 400,000 kw of generating capacity on one bus at Coulee Dam, a fault on or adjacent to this bus is a major consideration, and will result most surely in a major out-of-step condition on the rest of the system if not cleared promptly. For this reason the fastest circuit breakers available (3-cycle interrupting time) combined with the fastest system of relaying (carrier relaying) is used on the Bonneville System and at Grand Coulee to insure minimum disturbance and maximum stability.

The 230-Kv 3-Cycle Oil Circuit Breakers

The principal design features of the 230-kv 3-cycle oil circuit breakers of the type used for the fault tests already have been described.[3] Figure 4 shows a photograph of the test breaker *JYA* which is located at the east end of the 230-kv switchyard and is used for switching one of the lines to Spokane.

The housing on the side of the first tank encloses a compressed air operating

mechanism with tripfree mechanical linkage which is capable of closing the breaker in approximately one-third second using an air pressure of 275 pounds per square inch. The high speed triple latch assembly permits the breaker to part contacts within $1^1/_2$ cycles after energizing the trip coil.

In order to obtain the maximum short-circuit current at the instant of contact parting during this particular series of tests, a special tripping switch was attached to the mechanism frame in such a position that the main lever would close the switch contacts about a cycle before the breaker contacts touched during the closing stroke. This switch can be seen in the close-up view of the mechanism in Figure 5. Thus, when closing against a fault, tripping could be initiated sufficiently in advance of the establishment of the short circuit that the contacts would separate during the first half cycle of fault current.

Each phase of the breaker is built in a steel tank, 108 inches in diameter, with capacitor type bushings having potential taps which were used to obtain an oscillographic record of the transient voltages during the breaker operation. The cross section view of Figure 6 through one of the breaker pole units shows the linkage for operating the main lift rod and cross-arm bridging the two interrupting units on the ends of the capacitor bushings.

A sectional view of the multiflow deion grid type of interrupter used in these 3,500,000-kva circuit breakers is shown in Figure 7. The effectiveness of this device is indicated by the fact that only one pressure generating break and one interrupting break per breaker terminal is required for 3-cycle operation at 230 kv. Gas pressure developed in the upper chamber by the short-circuit current sends a flow of oil down two parallel channels and then through pairs of inlets in order to converge on the main arc from opposite directions at four different levels. The flow sweeps axially upward and downward through orifice openings to the vent passages interspersed between the inlets, cooling and deionizing the arc space so that interruption takes place at an early current zero. Arc energy is kept low by avoiding unnecessary lengthening of the arc.

To supplement the self-generated flow during the interruption of line charging current, a spring driven piston is located in the bottom of the contact assembly in such a way as to operate independently of the movement of the lower contact. Oil flow from this piston passes through channels to a single orifice in the top pressure gap and then down the main channels to the interrupting break. During high current interruptions, back pressure holds the piston until after arc extinction. As the pressure falls the piston then completes its stroke, sending a flushing flow of oil through the structure.

A view of two of these multiflow deion grids taken out of the test breaker is illustrated by Figure 8. The shields have been removed to show the contact operating linkage and the vent openings in the side of the grid assembly.

Line Dropping Tests

A series of line dropping tests was made with breaker *JY*-31 opening the charging current of the 183 miles of line from Coulee Dam through Columbia substation to Covington near Seattle. On some of the tests there were single delayed arc reignitions, but the majority of operations were without any restrikes as illustrated by Figure 9, showing complete opening 2.8 cycles after the breaker trip coil was energized.

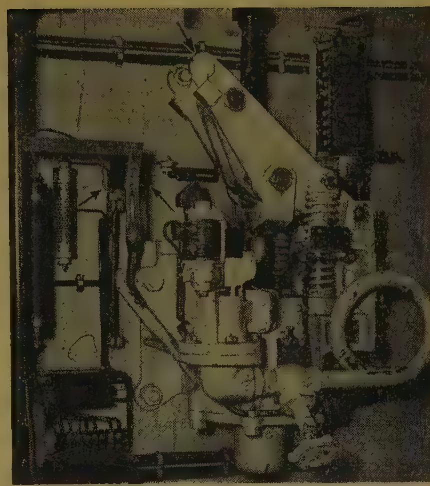

Figure 5. Compressed air operating mechanism for 230-kv test breaker

Arrows point to special pretripping switch

Of particular interest is the record of voltage E_a directly across the oil circuit breaker terminals. This was measured by means of an amplifier circuit fed from the voltage difference between the potential taps of the two capacitor type bushings on pole *A* of the test breaker. Note that the recovery voltage builds up to double peak value, and then oscillates about an axis displaced by an amount representing the trapped charge on the

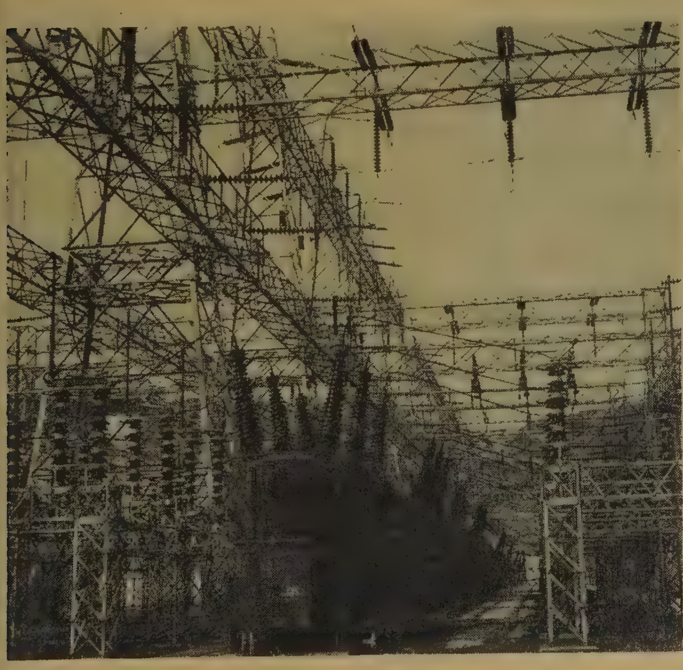

Figure 4 (left). View of Grand Coulee Dam switchyard showing 230-kv oil circuit breakers including test breaker *JYA*

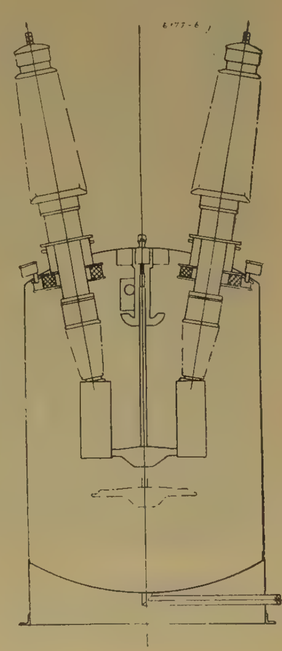

Figure 6 (right). Section through one pole of 230-kv 3-cycle breaker equipped with a pair of interrupting units rated at 3,500,000 kva

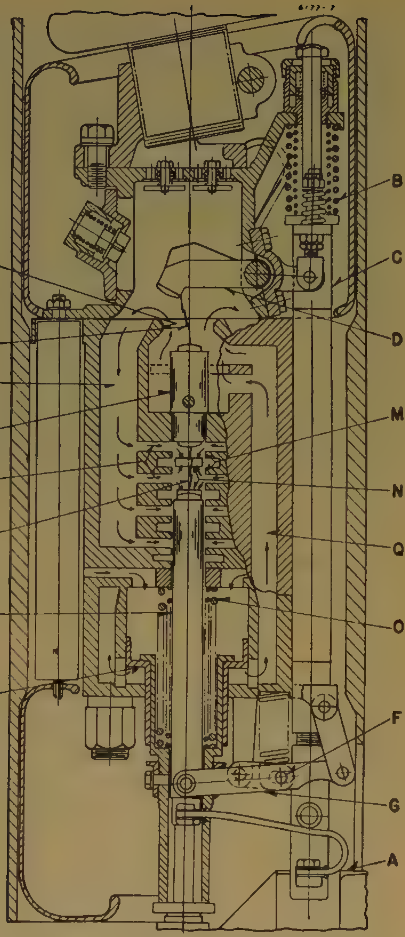

Figure 7. A 230-kv multiflow deion grid interrupting unit showing oil flow path from piston used to interrupt line charging current

Tripping of breaker releases main cross arm *A*, allowing springs *B* to move operating rod *C* down and rotate contact arm *D*, drawing pressure generating arc *E*. Simultaneously, pin *F* swings lever *G* downward, pulling contact *H* away from intermediate contact *I* to draw main arc *J*. Pressure from upper arc *E* drives oil through channels *K* to multiflow inlets *L*, deionizing arc in multiorifice structure *M*, and sending arc products out vents *N*. Springs *O* push piston *P* downward to provide supplementary oil flow for low current arc interruption, flowing oil through channel *Q* to orifice *R* in upper gap and then to main flow channel. Back pressure from high current arcs stops piston until arc is interrupted, when completion of stroke sends flushing oil flow through grid

line which very slowly leaks off. The records of line voltage E_b and E_c, being obtained with conventional potential devices including a step-down transformer, fail to show the sustained charge on the line after being disconnected from the Coulee bus.

Short-Circuit Tests

Single phase tests at 132 kv to ground have been made on a similar breaker in the high power laboratory up to 10,400 amperes, corresponding to a 3-phase fault of 4,300,000 kva. The rate of rise of recovery voltage on this freely oscillating laboratory circuit is approximately 2,500 volts per microsecond, representing a more severe operating condition than is ever likely to be encountered in service.

The 230-kv bus at Coulee Dam, when all transmission lines are disconnected, approximates the laboratory conditions, although it is not likely that maximum generating capacity would be connected to a bus section with only one line carrying load. However, there is more capacitance to ground on the Coulee bus than a laboratory circuit because of the greater number of connected pieces of equipment such as breakers, transformers, bus supports, and coupling capacitors. This is sufficient to reduce the natural frequency of oscillation at the bus, and likewise the voltage recovery rate, to less than 40 per cent of the laboratory circuit.

It should be pointed out that although test breaker *JYA* has been in service for about four years, it was revamped with the present contacts for 3,500,000 kva rupturing capacity about a year and a half before these special tests.

Table I. Fault Tests—Oil Circuit Breaker *JYA*

December 2, 1945

Generating Units *L*-2, *L*-4, *L*-5, *L*-6, *L*-7, *L*-8*

Duty Cycle	Test Number	Oscillogram Number	Type of Fault	Spokane Number 4 Line	Bus Voltage, Kv	Phase	Instrument Current Transformer	Bushing Current Transformer	Fault Current,‡ RMS Amps	Interrupted Kva on 3-Phase Basis	Interrupting Time, Cycles
CO-15-CO	1	35	3 phase to ground	On	241.5	A	5,000/1		6,908	2,890,000	
						A		300/1	7,000	2,929,000	2.7
						B		300/1	7,425	3,106,000	2.5
						C		300/1	6,240	2,610,000	2.8
	2	36	3 phase to ground	On	241.5	A	5,000/1		5,900	2,470,000	
						A		300/1	6,350	2,658,000	2.4
						B		300/1	5,940	2,485,000	2.7
						C		300/1	6,630	2,770,000	2.6
CO-15-CO	1A	40	Phase A to ground	On	241.5	A	5,000/1		6,360	2,660,000	2.5
	1B	41	Phase A to ground	On	241.5	A	5,000/1		7,840	3,280,000	2.4
CO-15-CO	2A	42	Phase C to ground	Off	241.5	C	5,000/1		10,450	4,370,000	2.7
						C		300/1	†8,866	†3,708,000	
	2B	43	Phase C to ground	Off	241.5	C	5,000/1	...	10,450	4,370,000	2.7
								300/1	†5,006	†2,094,000	
CO-15-CO	3A	44	Phase B to ground	On	241.5	B	5,000/1		10,015	4,190,000	2.8
								300/1	†7,211	†3,016,000	
	3B	45	Phase B to ground	On	241.5	B	5,000/1		6,535	2,735,000	2.7
								300/1	†6,677	†2,793,000	

* See Figure 10 for bussing arrangement.

† These current values are not reliable because of saturation of the 300/1-ampere bushing current transformers. The 5,000/1-ampere instrument current transformer has a 5,000-ampere primary rating, with a 1-ampere secondary.

‡ These currents are not necessarily the maximum fault currents obtained, but are the currents existing at the instant of mechanical separation of the breaker contacts, upon which the interrupted kva is based.

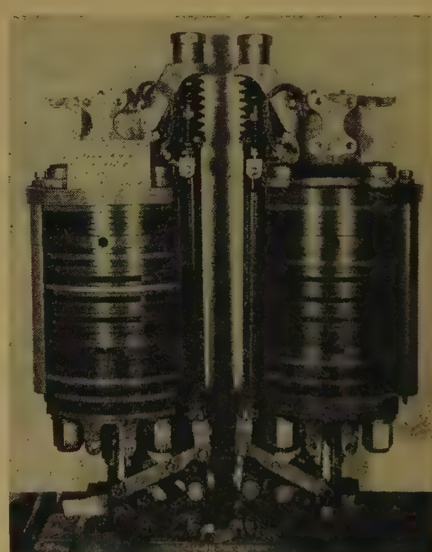

Figure 8. View of pair of 230-kv multiflow deion grid units with shields removed taken after high power fault tests

Three-Phase Fault Tests

Following two preliminary tests with one generator and with three generators connected, a standard duty cycle test of two close-open operations separated by a 15-second interval was made under a 3-phase grounded-fault condition with six generators connected. These included four 108,000-kva units and the two 75,000-kva generators as shown in Figure 10. The maximum power interrupted, as indicated by the data of Table I for tests 1 and 2, was 3,100,000 kva, approximately 89 per cent of the breaker rating. The breaker was pretripped as described, permitting the contacts to be separated during the first cycle of short-circuit

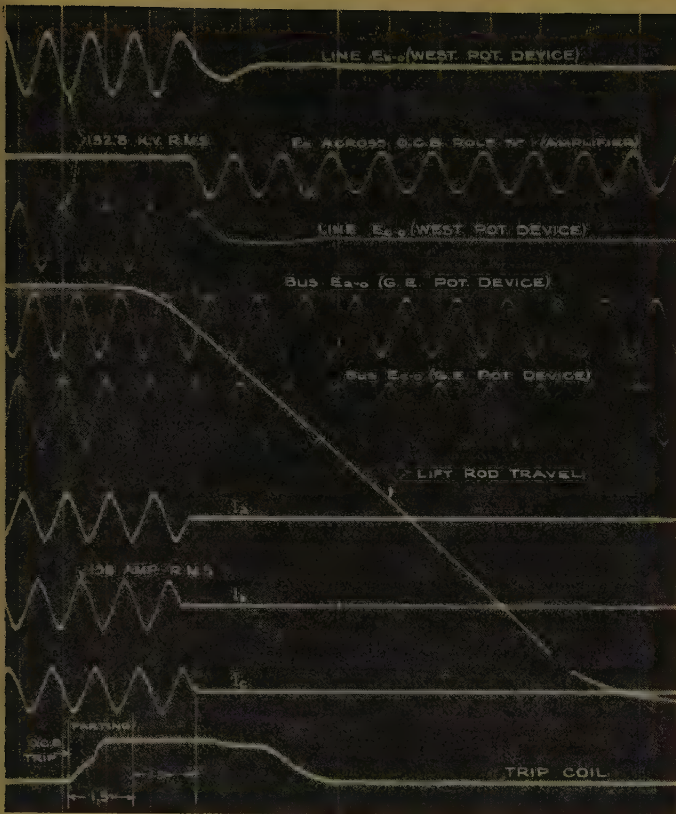

Figure 9. Oscillogram of interruption of charging current to 183 miles of 230-kv line

Record E_a through amplifier shows voltage across pole 1 displaced by charge on line

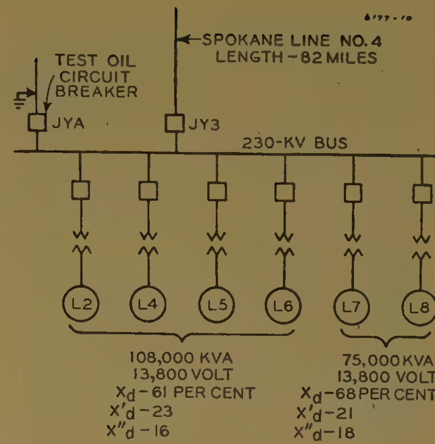

Figure 10. Diagram of connections for short-circuit tests on 230-kv 3-cycle breaker *JYA* using four 108,000-kva generators and two 75,000-kva units in parallel

current (see Figure 11). The performance of the breaker was remarkably quiet, no smoke or jar being evident to the observers.

Before proceeding with the single-phase short circuits at current values up to and above the breaker rating, the oil was removed and an inspection made of the condition of the contacts. So little material was burned away that it was not considered necessary even to smooth up the contact surfaces or make any changes in the contact adjustments. The oil and insulating surfaces were quite clean, so that the breaker was refilled immediately for the remaining tests.

Single-Phase Fault Tests

Each pole of the breaker was given the standard duty-cycle test consisting of two close-open operations separated by a 15-second interval, with a single line-to-ground fault close to the breaker and six generators feeding power to the bus. The pretripping scheme was used again to obtain contact parting as close to the first cycle as possible. Data from the oscillograms of these six interruptions also are given in Table I.

Tests 2*A* and 2*B* form a remarkable pair of almost duplicate interruptions of 10,450 amperes at 241.5 kv bus voltage, corresponding to 4,370,000 kva on a 3-phase basis. The breaker trip coil was energized one cycle before the closing contacts established the fault current which consisted of only three half waves, two large almost completely displaced loops with a small reverse loop between (see Figure 12). The breaker performed quite easily on all of these high power operations, there being very little external evidence of the tremendous short-circuit power cleared by the switch. Since no transmission lines were connected during these two tests, the bus voltage was permitted to oscillate freely, giving a voltage recovery rate estimated from the oscillograms to be close to 1,000 volts per microsecond.

It is interesting to compare the short-circuit current values actually measured with the calculated figures. The maximum asymmetrical 3-phase short-circuit power from four 108,000-kva and two

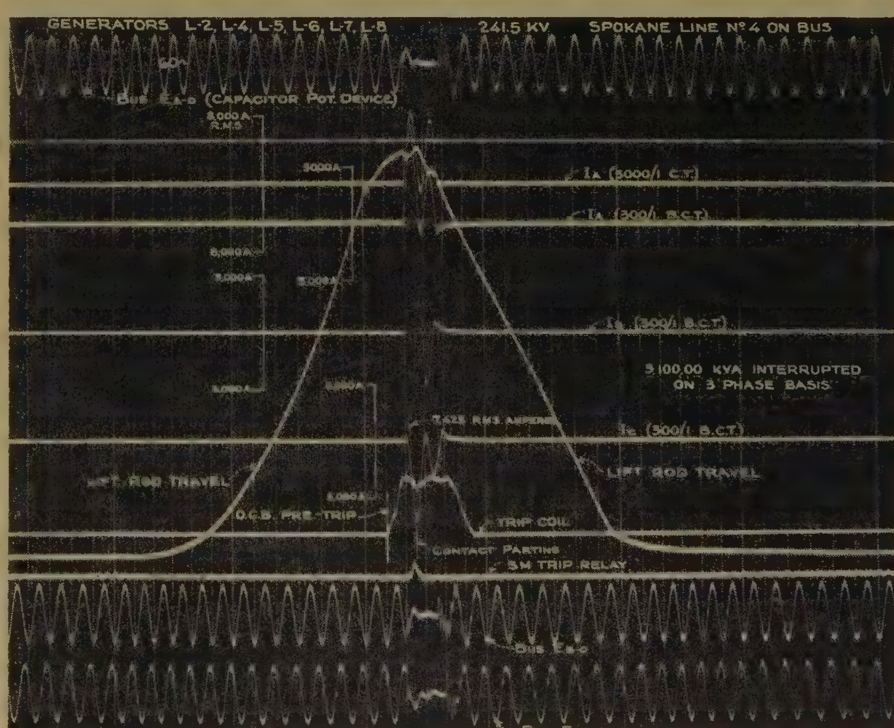

Figure 11. Oscillogram of 3-phase short-circuit test at 241.5 kv interrupting 3,100,000 kva

Breaker pretripped one cycle before fault established by closing contacts

75,000-kva generating units in parallel on the 230-kv bus, with two cycles of decrement, amounts to 2,970,000 kva from Hoard's data.[1] The curve in Figure 13 shows a decrement factor of 1.4 at two cycles, and 1.64 at one half cycle where the breaker contacts actually parted and the current was measured on these tests. Multiplying by the ratio of these two factors raises the short-circuit power to 3,480,000 kva, corresponding to a maximum asymmetrical first-half-cycle current of 8,310 amperes with the power transformers on the 241.5-kv taps used for the tests.

For single-phase faults, with all of the transformer neutrals solidly grounded, the current should be at least 30 per cent higher or 10,800 amperes, which is approximately the maximum current obtained.

On the other four interruptions, the same six generators were connected, but varying degrees of asymmetry in the current wave resulted in short-circuit power between 2,500,000 and 4,000,000 kva on a 3-phase basis. For these tests Spokane number 4 line, 82 miles long, was connected to the high voltage bus, cutting the voltage recovery rate to about a third and damping out overswinging of the initial transient. No consistent effect on the interrupting performance was observed, however, because of this difference in recovery rate, the interrupting times on the six high-power single phase tests being included in the narrow range from 2.4 to 2.8 cycles with an average of 2.65 cycles.

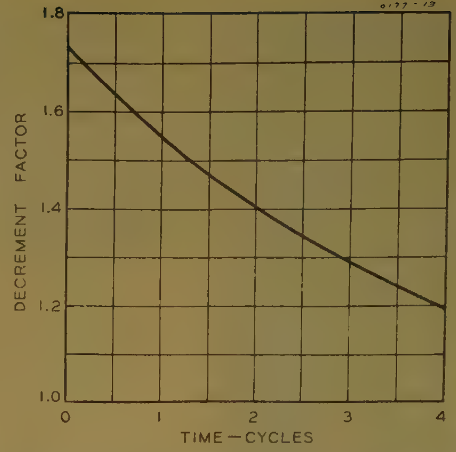

Figure 13. Calculated decrement of maximum asymmetrical fault current for short circuit at Grand Coulee 230-kv bus

Test values checked this curve very closely

The oil dielectric strength after this heavy duty measured as high as before the tests, and there was only a slight discoloration caused by carbon formation from the arcing. All of the multiflow grid surfaces were quite clean, with no appreciable enlargement of the orifice dimensions as indicated by Figure 14. The silver-tungsten contact tips had lost not more than $^1/_{32}$ inch of material over only a part of the surface area, and were still in condition for further service (see Figure 15). The only result of applying these heavy short circuits up to nearly 25 per cent above rating was the cracking off of a small sliver from one of the fiber plate items, and a dislodgement of one intermediate contact sufficient to keep it from falling back into its normal position. With minor design changes to increase the mechanical ruggedness still further there would seem to be no reason why this type of interrupter could not handle satisfactorily short circuits of 5,000,000 kva.

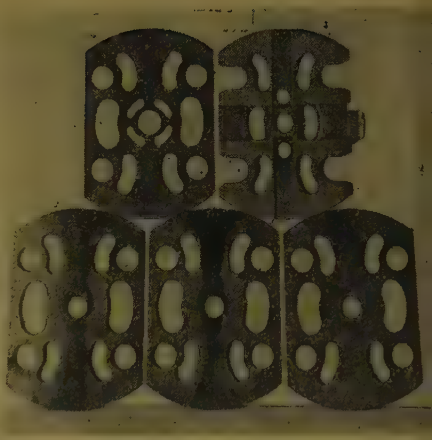

Figure 14. Photograph of fiber grid plates removed from multiflow grid after 4,370,000-kva test showing only slight erosion of orifices

Small chip of fiber was split off upper right plate

Figure 15. Photograph of arcing contacts from pole 2 of 230-kv breaker after 4,370,000-kva test showing only moderate burning of silver-tungsten tips

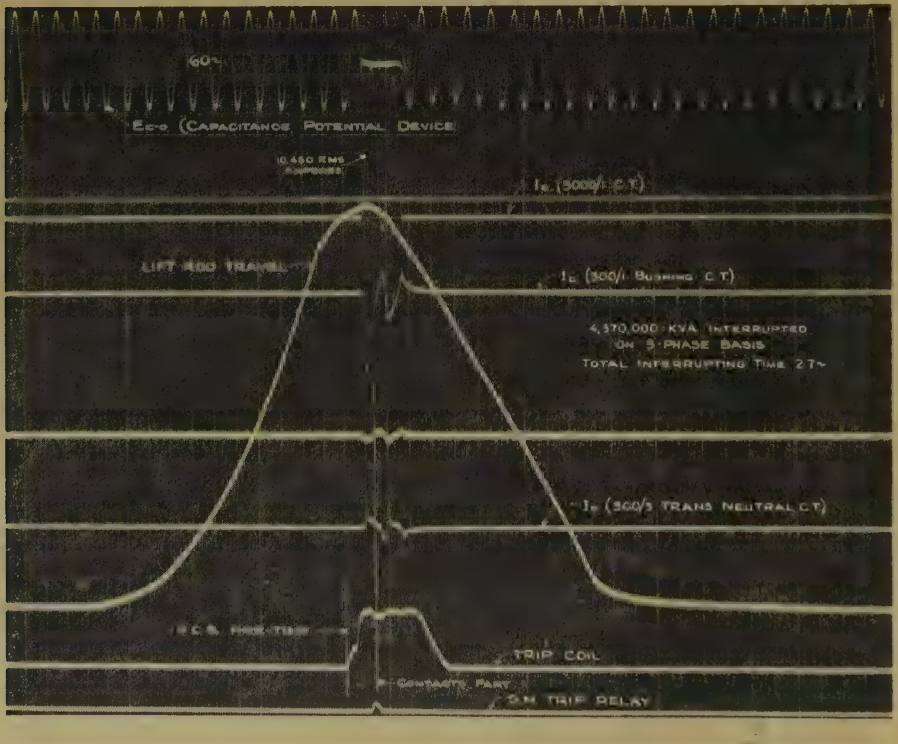

Figure 12. Oscillogram of maximum power single-phase fault interruption, equivalent to 4,370,000 kva at 241.5 kv 3-phase, within 2.7 cycles after energizing breaker trip

No lines connected to bus

Conclusions

These heavy short-circuit tests above 4,000,000 kva on the 230-kv bus at Coulee Dam have served not only as a check on calculations of fault currents, but they also have confirmed high power laboratory test in demonstrating a margin of safety in the rupturing capacity of the 230-kv 3-cycle breakers as rebuilt for a rating of 3,500,000 kva. The relative ease with which the test breaker performed its duty suggests the practica

Philosophy of Relaying

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PHILOSOPHICALLY the electric power system exists to serve certain human needs and it must serve them better and more economically than they could be served by other methods or it would not be able to continue to exist or to grow. The rendering of service of the required degree of excellence, within economic limitations, is not an inherent attribute of the electric power system but is the result of sustained and carefully co-ordinated effort on the part of all concerned in the rendering of the service, including

1. Precautions taken in the design of equipment to make it trouble free and reliable.

2. Precautions taken in system layout and connections to enable service to be rendered continuously under adverse as well as under favorable conditions.

3. Precautions taken in the operation and maintenance of the system to avoid unnecessary interruptions to service.

When everything economically possible has been done along these lines, however, the system still remains subject to unavoidable defects which result in faults, tending to cause damage to equipment and interruptions to service. Specific measures taken to minimize the detrimental effects of such faults comprise system protection, the fundamental idea of which is to reduce damage to equipment and to improve service to customers. Relay protection cannot prevent faults from occurring on the power system, but it is a means of minimizing the effects of such faults when they do occur. The expenditure which can be justified for protection is a measure of the probable reduction in cost of repairs and of the probable improvement to service which will result from the use of the protection.

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General Principles of Protection

The basic ideas and principles which underlie and determine the policies adhered to in providing protection for the electric power system are relatively few and simple. They may be divided conveniently into two groups.

The immediate object of the electric power system is to make continuously available a supply of energy at approximately constant voltage and frequency. A difficulty encountered in attaining this objective results from the fact that a fault at one point on the system tends not only to destroy or seriously damage the equipment directly affected but also to involve the whole system in serious voltage and frequency disturbances and possibly in widespread interruptions to service; the seriousness of these effects being, in large measure, determined by the length of time that the fault is allowed to remain on the system. This difficulty is met by building the system up of numerous elementary sections joined together at strategic points by suitable automatic circuit-interrupting devices which enable any section to be isolated from the remaining sections without interfering seriously with the essential integrity of the system as a whole. Each of these sections is then provided with appropriate automatic equipment designed to detect faults within the section and to quickly and selectively isolate a defective section from the remainder of the system. In order that no part of the system may remain unprotected the various zones of protection are made to overlap each other at the junction points of the respective sections.

Not only is system protection concerned with the successful isolation of defective sections of the system, it is concerned equally with the determination of fault locations and causes and with the return to service of affected sections in as short a time as possible. For this purpose, therefore, extensive use is made of relay indicators, automatic fault locators, and automatic oscillographs in the analysis of system troubles. Furthermore, experience has shown that the majority of faults on transmission circuits prove to be transient if they are cleared quickly. In such cases automatic reclosing breakers are used as an additional means of improving service.

From the foregoing it may be concluded that, to a very large extent, the protection of a large power system must be a custom built job. The system is analyzed carefully and whenever major modifications or additions take place the necessary fault studies are carried out. As a result of such studies appropriate protective schemes are applied. In general, these schemes are built up in the field, of standardized units of equipment, assembled into the various special combinations necessary to meet the requirements of any situation that may be encountered.

Another fairly obvious conclusion is that, in order to obtain the degree of coordination required for successful functioning, the protection of the whole system must be placed under the control of one centralized authority.

Applications of Principles

When attempting to apply the principles of protection it is necessary to note the physical conditions that can be made use of as the basis of the automatic features of protection. Of primary importance in this respect are the two active quantities of the electric circuit—current and voltage. Intelligently used in appropriate circuits and devices they can be made to represent most of the characteristics that are essential in distinguishing between normal and abnormal conditions on the system. To a more limited extent, various other physical conditions associated with electrical apparatus such as temperature, pressure, gas accumulation, and frequency, may be used also. In ad-

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bility of constructing breakers of still higher interrupting rating. The application of such breakers at Coulee Dam would free the system from present operating limitations.

The success of these record breaking tests is the result in no small measure of the excellent co-operation of many members of the engineering and operating personnel of the Bureau of Reclamation, Bonneville Power Administration, and Westinghouse Electric Corporation.

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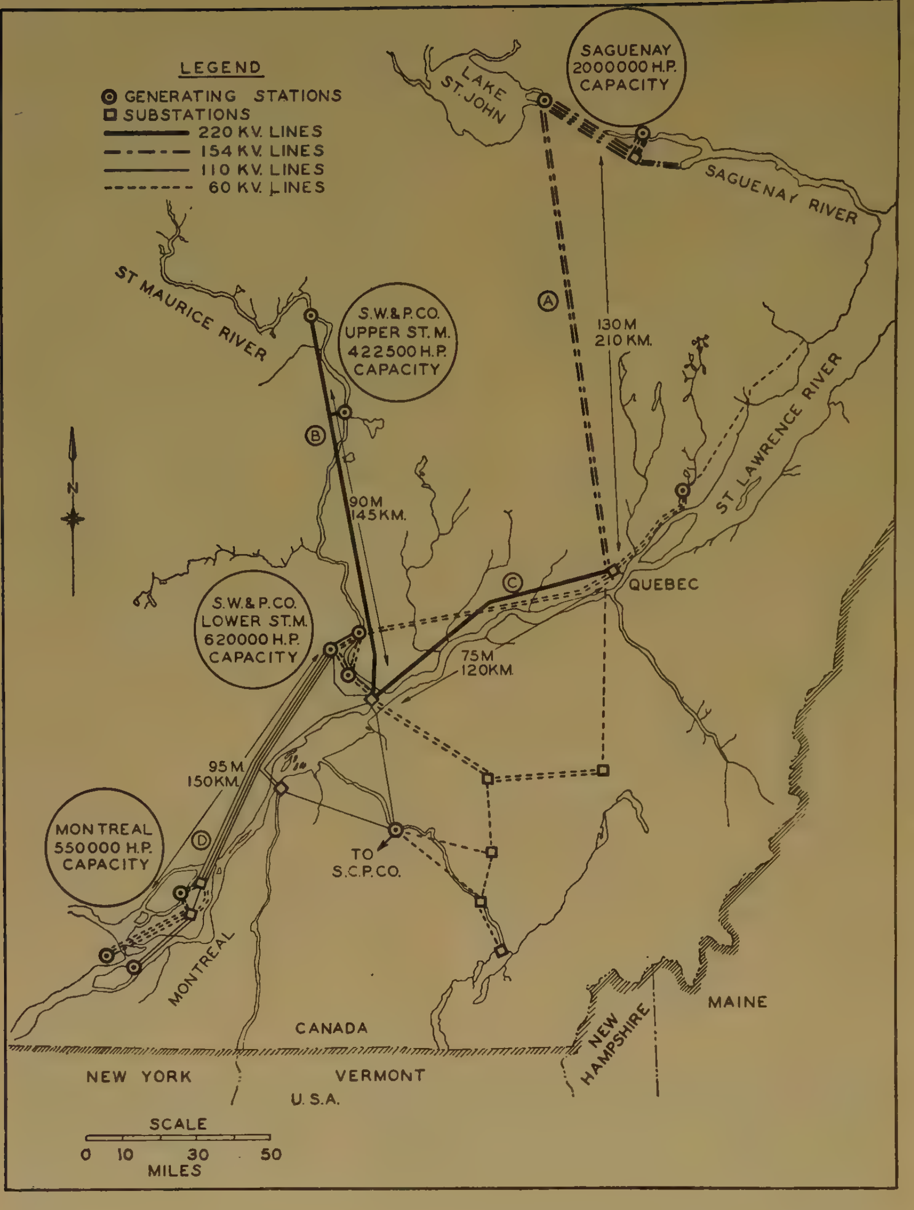

Figure 1. Map showing main transmission lines of interconnected systems, Province of Quebec

dition to these, time is quite generally useful as a co-ordinating factor.

Protection may be considered to be simple or complex depending on whether it makes simultaneous use of one or more of the fundamental quantities available for fault detection. Furthermore, complex protections may possess varying degrees of complexity.

The simplest form of protection is the fuse which utilizes current only and finds extensive applications on a great variety of electric circuits from the lowest voltages up to 138,000 volts. It incorporates in itself both the fault detecting and circuit-interrupting functions. Its great advantages are low cost and simplicity, but its use is limited to circuits carrying a few hundred amperes. An outstanding disadvantage is the more or less prolonged interruption to service required for replacement whenever there is an operation. This may be overcome to some extent by application of the principle of the reclosing fuse. Their somewhat inflexible time–current characteristics may at times be another disadvantage of fuses.

Some of the limitations of fuses may be overcome by utilizing electromagnetic coils energized by the current flowing in the protected circuit. It is then necessary to provide a separate circuit-interrupting device, or circuit breaker, which the coils may actuate. The simplest form of protection associated with a circuit breaker consists of a trip coil on the breaker actuated by the current, or a proportion of the current, which passes through the breaker. Such trip coils may be provided with dash pots to give them an adjustable inverse time–current characteristic. A means is then available for obtaining selective action between circuits on a system so protected. The flexibility of this form of protection may be increased by separating the adjustment and timing features from the trip coil and incorporating them in relays, leaving the trip coil the sole duty of releasing the latch on the circuit breaker.

In these simple forms of current protection there are only two possibilites of obtaining selectivity. One is the possibility that the relative magnitudes of fault currents, at the points between which selectivity is required, will differ by a sufficient amount to inherently provide the required selectivity. The other is the possibility of using time to provide the required selectivity. The former possibility is usually a forlorn hope. The great disadvantage of the latter is the delay involved in clearing faults, which may become several seconds on an extensive system. Another disadvantage is the fact that, in obtaining selectivity by timing, the shortest clearing times are at the ends of branch feeders where fault currents are minimum whereas, from the point of view of limiting damage, the shortest clearing times should be nearest to the power source where fault currents are maximum. Even accepting the kind of selectivity that can be obtained by the use of time and simple current, the method is applicable to radial systems only and not at all to networks.

The field of usefulness of protective features based on current only may be extended very appreciably by associating with it the principle of comparison. This gives rise to such schemes as:

1. Differential current which may be applied to transformers, generators, busses, and short tie lines. Advantages of differential current protections are the fact that the zones protected are clearly and sharply defined providing inherent selectivity; settings may be very sensitive and time instantaneous. Such protections involve problems of current transformer characteristics.

2. Current balance protections such as parallel line and generator split phase. An advantage of the split phase protection is that it can be made very sensitive. Its zone of effectiveness is also clearly defined. Parallel line protections are ineffective for line end faults. On radial lines the zone of ineffectiveness may be a considerable percentage of the line length.

3. Phase sequence networks actuated by current only may be used in conjunction with pilot wires to provide current comparison between the two ends of a transmission line, giving the equivalent of a differential protection.

4. Another method of protecting transmission lines by means of current only is to utilize phase sequence networks in conjunction with carrier current to compare the relative directions of current flow at the two ends of the line.[1]

Also included in the simple protections are

1. Overvoltage, used on generators.
2. Residual voltage, for generator neutrals.
3. Undervoltage.
4. Unbalanced voltage.
5. Gas accumulation, used on transformers.
6. Pressure, used on transformers.
7. Temperature, used on various kinds of equipment.
8. Overfrequency.
9. Underfrequency.

Among the complex protections, impedance is probably the simplest. In principle it uses current and voltage to measure the impedance to a point of fault. When the impedance drops below a critical value operation results. It has the advantage that its distance range, or zone of effectiveness, may be limited at will, within reason. It has the further advantage that it is most sensitive for near-by faults and increasingly insensitive as the end of its zone of effectiveness is approached. Its field of usefulness is in the protection of lines, especially on networks, and for standby or backup protection. It has the disadvantage that its range is altered by power arc voltage and by fault resistance. By associating it with a single time-step effective co-ordination of the protection on a radial system may be obtained without an inordinate increase in time of clearing faults at any point. On network systems it may be associated with directional features of protection to attain an effective co-ordination of the protection of the whole network. In cases where changes of generator capacity connected to the system would affect a simple current protection this limitation is not found with impedance protection.

The disadvantage of having the distance range of an impedance relay affected by fault resistance can be overcome by using a reactance relay. If necessary to attain selectivity a directional impedance or reactance protection may be combined with pilot features by using direct current on metallic conductors or by using carrier on the high voltage line. Under this condition the distance range adjustment need not be critical and simultaneous clearing of both line ends is obtained for all line faults.

On long heavily loaded lines impedance relays may get into trouble by tripping on overload or on power swings. Such a condition may be improved by the use of relays which measure impedance at an angle.

When sections of a large power system tend to pull out of synchronism due to clearing faults, power surges over certain tie lines may cause unnecessary tripping. In such cases when there is the probability that the swings would subside and conditions return to normal, out-of-step blocking may be added to protections that are sensitive to these conditions, to prevent faulty tripping. In cases where it is probable that synchronism cannot be maintained after a swing starts, conditions can sometimes be improved by separating the system at a selected point. For this purpose a system separator can be added to the appropriate protection.

The protective features discussed thus far have been those designed to detect and selectively isolate faults. There remains to be considered, methods of automatically returning circuits to service following the isolation of a fault. In general automatic reclosing is not applicable in the case of equipment failures because defective equipment requires inspection and repair before it can be returned to service. On lines and feeders however the majority of faults are transient and the circuit will remain in service if reclosed immediately. With high speed protection should reclosing take place onto a permanent fault no particular harm would result.

A successful scheme of automatic reclosing can be based on the principle that if a fault is transient the circuit will stay in if it is reclosed immediately after it has cleared. If a fault is permanent the circuit will trip again immediately after it has been reclosed and no further attempts at reclosing should be made. There does not appear to be any practical use in providing more than a single shot recloser. In practice such a recloser is built to lockout should the circuit to which it is applied trip twice within a predetermined time interval, which may be anything from about 2 to 15 or more seconds.

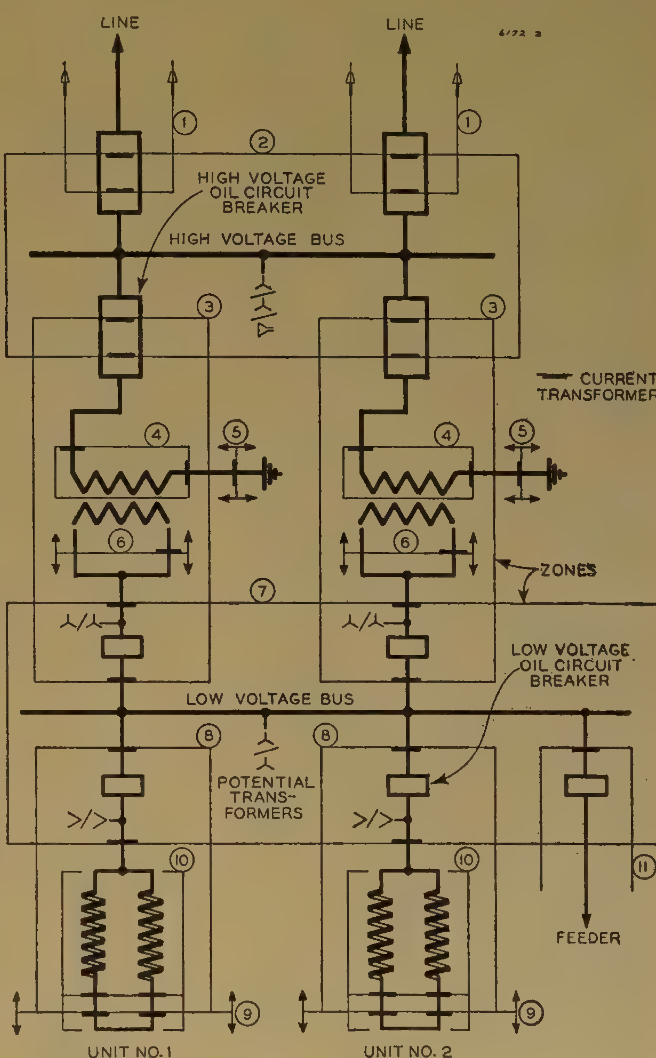

Figure 2. Diagram showing how a power system is covered by overlapping zones of relay protection

Number	Protection
1	Line
2	High voltage bus
3	Transformer differential
4	Transformer high voltage ground differential
5	Transformer high voltage ground stand-by
6	Transformer stand-by
7	Low voltage bus
8	Generator differential
9	Generator stand-by
10	Generator split phase
11	Feeder

Circuit breaker duty ratings have a bearing on automatic reclosing. If circuit breakers have ample capacity the time allowed for resetting the recloser may be short, if not the time must be longer.

Another factor which must be considered in providing automatic reclosing is the condition of synchronism. If a line is a tie between systems it must be decided whether to permit reclosing out-of-phase to take place, depending on the systems to pull into step, or to provide an automatic synchronizer, to operate in conjunction with the automatic recloser.

Practical Aspects of the Art of Protection

Up to this point the problems of protection have been dealt with in broad outlines only. Something now will be said about technical details, remarks along this line being in accord with current practice on the system of the Shawinigan Water and Power Company.

This system is located in the Province of Quebec and serves an area of roughly 30,000 square miles. It comprises a number of hydroelectric generating stations, having a total installed generator capacity of 830,000 kw. There are more than 60 high-voltage lines operating at 44 kv and above, totalling approximately 2,000 miles; in addition to an extensive network of low-voltage distribution lines operating at 33 kv and below which are operated by the distribution department. Much of the power is utilized in heavy industry such as paper mills and mining and electrochemical plants, in addition to that supplied to distribution centers. Interconnections (see Figure 1) are provided with the Saguenay, Hydro-Quebec, and other power companies, comprising a total of approximately 2,500,000 kw of generator capacity for the combined 60-cycle interconnected system.[2]

From the standpoint of protection this extensive system is composed of numerous elements joined together by means of automatic circuit breakers and, to a minor extent, by means of fuses. The various elements of the system are, in the aggregate, provided with a multiplicity of protective schemes which require the use of approximately 10,000 relays of 300 types. As far as possible each element of the system is provided with a protection, the zone of effectiveness of which overlaps the protected zones of adjacent elements. For convenience, relay protection zone diagrams are provided for the various sections of the system. Figure 2 is such a diagram for a small section of the system. This diagram depicts all the main elements of the system, namely, generators, transformers, busses, lines, and feeders; and may be used conveniently as the basis of a discussion of some practical details of protection design.

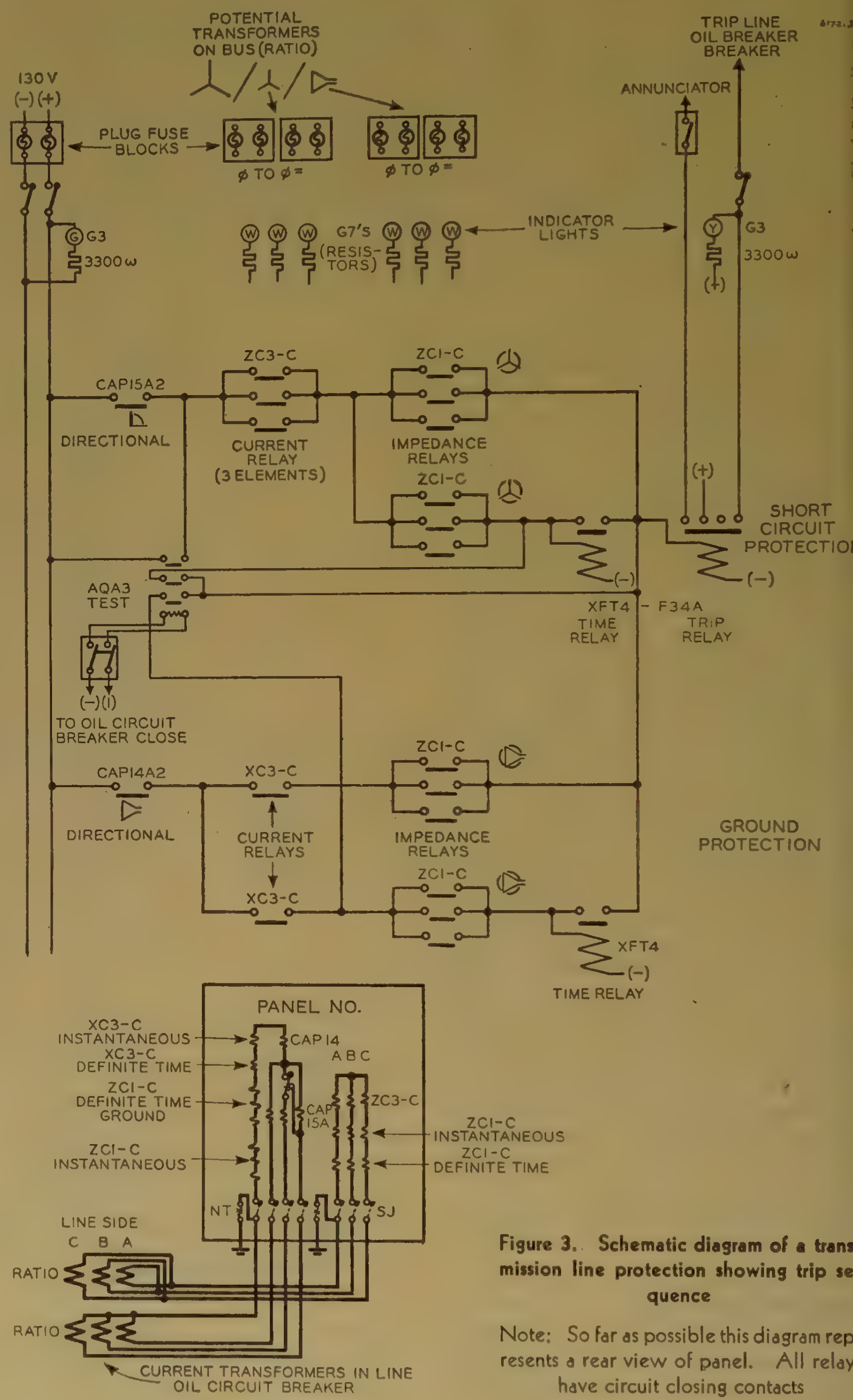

Figure 3. Schematic diagram of a transmission line protection showing trip sequence

Note: So far as possible this diagram represents a rear view of panel. All relays have circuit closing contacts

constructed during the past 20 years the tendency is to operate a unit system, each generator and its associated power transformer forming the unit. In some stations transformer banks are not provided with high voltage circuit breakers. High voltage lines, also, sometimes are not provided with high voltage circuit breakers. In all these cases the best expedients available are adopted to make the protection as effective as circum-

The hypothetical section of system shown in Figure 2 is more liberally provided with circuit breakers than is often the case in practice. In some of the older generating stations a main low voltage bus has been provided, but in stations

stances permit. One of these expedients is the high voltage fuse for transformer protection.

In general high voltage fuses are used on power distribution transformers and on potential transformers. It was formerly the practice to install glass-enclosed lead fuses on small and medium capacity transformers. In case of faults above the rupturing capacity of these fuses, the fault would be transferred to the high voltage line, causing it to open automatically by the action of its protection. Recent practice is to provide fuses of suitable timing characteristics and ample rupturing capacity for the respective location. Both dry and liquid types of fuses are used but preference is now given to the dry types because of their greater convenience and economy in use. Timing characteristics include: high speed fuses for potential transformers and for distribution transformers where selectivity is not required or where selectivity requires fast clearing; medium speed fuses where required for selectivity with other fuses or very fast circuit breakers; and slow acting fuses where selectivity is required with slow operating circuit breakers and relay protection. Fuses should only be used where it is not economically possible to justify the installation of an adequate relay protection.

Where high voltage fuses do not provide an adequate protection and where it is not possible to justify the installation of an automatic circuit breaker automatic grounding switches are used. The operation of the relay protection initiates the closing of the ground switch and this in turn permits the respective main line to be cleared by the action of its protection. Single-pole, 2-pole, and 3-pole switches are in use, either spring operated or motor operated. Experience has indicated that the motor mechanism is more reliable than the spring and that a 2-pole switch is usually adequate.

Protective relays are energized in general by current and potential transformers. Recent practice has been to utilize, as far as possible, bus potential transformers for line relay protection up to and including 110 kv. On systems operating at voltages above 110 kv either capacitor-type potential devices or compensated low voltage potential transformers are used for line relaying. Potential transformers are commonly provided with double secondaries to facilitate obtaining a delta point voltage for ground protection when required. When such a voltage is required and the potential transformers have only one secondary a set of one-to-one ratio auxiliary potential transformers may furnish the required voltage. Occasions also may arise where auxiliary potential transformers furnish a convenient means of correcting phase angle or ratio in the potential supply.

Current transformers quite commonly define the limits of protective zones. This is particularly true in the case of differential protections, but an exception may be noted where a feeder carrying a small load is permitted as a leak in the differential zone. Various justifications may be found for this practice, such as lack of suitable circuit breakers. In the case of directional distance line protections the directional relays permit the current transformers to mark one limit of the zone of effectiveness of the protection, the other limit being determined by the setting applied to the distance relays. On pilot protections current transformers mark both limits of the protected zone.

The most extensively used current transformers are through types installed on the bushings of circuit breakers and transformer banks, or on cables and bus bars when voltages do not exceed 15 kv. Wound primary current transformers are used to some extent on low voltage circuits and on circuit breakers which do not lend themselves to the use of bushing type transformers. Some circuit breaker designs do not permit the use of built-in current transformers and separate oil-filled current transformers are used. In all cases an attempt is made to keep the resistance of the secondary windings of current transformers low, to keep lead burdens low and to use the maximum cross section of iron consistent with space limitations.

In protection design attention is first given to the system diagram, to determine what may be necessary or possible in the way of zones of protection. Then fault studies assist in the co-ordination of the various features of the complete protective scheme. Following this a trip sequence diagram is prepared for each protection, to serve as the basis for the preparation of the main wiring diagrams.

Generators

Generators of 10,000 kva capacity or above usually are provided with several features of protection, including

1. Instantaneous split phase.
2. Instantaneous current differential.
3. Definite time impedance standby.
4. Definite time or inverse time overvoltage.
5. Definite time or instantaneous ground alarm.

The ground alarm is usually provided through the medium of a single potential transformer in the generator neutral or through the installation of a set of star/corner delta potential transformers connected to the generator terminal leads. Thirty per cent of all generators are grounded through a single potential transformer.

In future installations serious consideration will be given to the installation of a small single phase distribution transformer between generator neutral and ground. Differential and split phase protections are set as sensitively as possible to reduce burning of iron laminations and thus avoid restacking. Settings of the order of two per cent of generator rating are attained. By-pass reactors on the relay coils assist in eliminating faulty relay actions due to d-c transients on through faults. Overvoltage protections are set to act instantaneously at about 140 per cent of normal voltage to avoid overstressing insulation following a sudden drop of load. The standby protection acts as a stand-by to the system protection at large and to the generator in case of the differential and split phase features are out of service.

Transformers

Power transformers are usually provided with one or more of the following forms of protection:

1. Inverse-time current differential or 2-step current differential. In some cases impedance relays are used instead of current relays for this feature.
2. Instantaneous residual current ground protection, connected differentially if it appears feasible or advantageous to do so.
3. Gas detector relay to indicate gas accumulation or pressure. In some cases the gas accumulation element is arranged to provide a visual and audible indication only, without actually tripping the respective bank.
4. Impedance stand-by, combined with generator stand-by on the unit system.

Large power transformers are usually provided with some form of differential protection, which either overlaps the generator protection, in the case of a unit system, or the low voltage and high voltage bus protections where busses are used. The aim is to make these protections as sensitive as possible to avoid damage to core iron and to reduce the hazard of oil fires.

In the 2-step form the sensitive feature is set at 30 per cent of bank rating

and 2.5 seconds definite time; the instantaneous feature is set at 225 per cent of bank rating.

Gas detector relays are used more particularly on large 3-phase power transformers and step regulators. Many 3-phase power transformers are equipped for tap changing under load and the gas detector relay provides a measure of protection for this device. Gas detector relays also are used on single phase transformers where they would be more convenient to apply than an adequate excess current or differential protection. Experience with these relays has been reasonably satisfactory and the gas detector element has operated for conditions such as low oil, air in the oil, stoppage of breathers, and overheated connections. Some difficulties have, however, been encountered because of inexperience of operating personnel, oil hammer when attempting to determine the presence of gas, and moisture collecting on the terminal studs of the relay.

Busses

Busses usually are equipped with some form of duplicate differential protection. This usually requires two current transformers per phase, unless a special arrangement is devised.

In some cases a scheme of double differential bus protection has been provided using a single set of current transformers with both impedance and current relays connected so that it requires both features to operate in order to trip the respective bus. Where fault currents exceed the interrupting capacity of low voltage circuit breakers an excess current feature may be added to the respective feeders to insure automatic clearance of the whole bus, rather than rely on the low interrupting capacity of individual circuit breakers. During recent extensions to the power system, it was estimated that fault current in the vicinity of one 110-kv station would exceed the interrupting capacity of existing circuit breakers. There were two 110-kv busses at this station and an arrangement was made to install a high speed circuit breaker between these two busses, equipped with a relay scheme to automatically open the bus tie circuit breaker for nearby faults. It is designed in this manner so that the busses will be separated before a local breaker attempts to clear a nearby fault, thereby reducing the fault current to a value which can be handled by the circuit breaker on the faulted circuit. This feature has operated on several occasions to date with satisfactory results.

Lines

Line protections vary greatly depending among other things on voltage, load, type of system, and type of line. In general the Shawinigan Water and Power Company favors the 2-step impedance and directional scheme for both short circuit and ground protections. The trip sequence diagram of such a protection is shown in Figure 3. This form of protection is particularly applicable to high voltage lines having low ground resistance and operated in either a parallel line or network system.

On lines having high ground resistance, such as wood pole lines, stand-by residual current relays are added to other features. In some cases use has been made of parallel ground relays for wood pole 60-kv lines operating in parallel at both ends. Some difficulty has been experienced in providing a fully satisfactory ground protection for such wood pole lines not equipped with overhead ground wires or other means of reducing ground fault resistance. On short tie lines connected into the main transmission network recent practice has been to provide buried counterpoise wires in order to improve the effectiveness of the impedance ground protections.

All power circuits of 110 kv and above are operated solidly grounded star. All 60-kv circuits are operated "delta isolated neutral" but with small grounding transformers located at strategic points. In some cases transformers of 1,000–3,000 kva capacity used for auxiliary supply are utilized also as grounding transformers. A circuit breaker frequently is installed in the neutral of the grounding transformer and arranged to trip automatically when grounds persist for a definite time of two to three seconds. Most of the high voltage steel-tower lines are provided with either overhead ground wires, buried counterpoise wires, or both. This tends to reduce lightning outages and to keep the ground resistance to a low value, permitting the successful use of impedance ground delays.

The instantaneous step of an impedance protection is ordinarily set to cover 95 per cent to 100 per cent of line length. The second step is set to reach into the next line section with a time delay of 0.5–0.8 second. With high speed circuit breakers and protections it is probable that selectivity can be retained with these time delays reduced to about 0.4 to 0.5 second. On 3-ended lines 3-step impedance protections have been used successfully.

Instantaneous protection over the whole length may be obtained by the use of pilot protection. Three forms are in use

1. A-c 3-wire system using low resistance conductor.
2. D-c 2-wire system using high resistance conductor.
3. Carrier current on the high voltage lines.

Outstanding disadvantages of the a-c pilot wire protection are its high cost limiting its use to very short tie lines and the fact that it does not provide inherently any stand-by or backup protection. The cost of carrier current protection limits its use to high voltage lines and to conditions where adequate protection cannot be obtained otherwise. The d-c pilot protection is used on lines up to 25 miles in length. It usually operates from dry cells of 90–135 volts rating. Conductors vary, depending on local conditions, and include private telephone circuits, leased Bell Telephone circuits, and small-conductor buried cable. All pilot cables are well insulated from ground. The d-c pilot scheme has the advantage that it can be added to a more or less standard line protection and will provide stand-by protection at all times, even when the pilot feature may be out of service.

Automatic Reclosing

There are several automatic reclosing installations in service on transmission systems operating at voltages up to 187 kv. New installations are being completed as rapidly as conveniently possible. Recent practice has been to adopt one immediate automatic reclosure with an automatic reset feature for transient faults and lock-out for permanent faults. Time delay of reclosure is normally kept to a minimum, usually not more than 2.5 seconds. To date, there are no installations of automatic reclosing on lines directly supplying synchronous motor loads. However, installations are now under consideration on a number of 60 kv lines supplying paper mill and mining loads, which include synchronous motor installations. An installation is also being completed on a system consisting of a small generator connected through a tie line to the main system, and on one of the radial lines connected to this there is a synchronous motor. We are looking forward with considerable interest to this development, with a view to extending automatic reclosing to network systems.

Conclusion

It will be realized that the application of protection to the power system does not

Control and Protection of Aircraft D-C Power Systems

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D-C POWER SYSTEMS on transport aircraft before World War II were small in capacity as compared to the present requirements. Present wide speed-range generators are 12 to 15 times the ratings used on old DC-3's, and system capacity on new transports may be 30 times greater and capable of sustaining fault currents greater than 5,000 amperes at 30 volts. It is also necessary to consider the increase in the voltage from the smaller system to the larger system. The 12-volt system does not have sufficient potential to maintain an arc or ionize gases so that an arc could be maintained for any length of time, whereas, the 24-volt system does have sufficient potential to maintain an arc or ionize gases. This is a fundamental characteristic of the electric circuit and is quite often forgotten about when voltages greater than the critical arcing values are used. Arcing voltages are in the neighborhood of 18 to 20 volts, depending upon the materials and atmospheric pressures at the electrodes involved. The number of electrically operated devices also has been increased, and many of the electrical requirements which are vital to flight operation are therefore dependent on the continuity of electric power. Study of various means of controlling and protecting electric power systems against failures in the light of these facts has led to some interesting simple developments which are described in this paper.

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Electric Power and Protective Systems

An aircraft d-c power system like any power system consists of generators, regulators, contactors, protective devices, functional relays, and cable. The simplest system consists of generators, regulators, and reverse current cutouts. This system on multiengine planes using 28 volts definitely has indicated that improvements are desirable from the standpoint of over-all performance, reliability, safety, and full automatic features, both during normal service and under conditions where faults cause serious consequences to the electric power system and perhaps to the airplane itself.

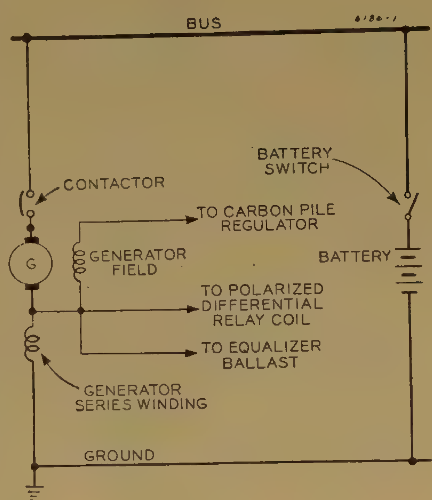

Figure 1. Diagram showing contactor at generator and no fault protection other than reversed currents and polarities

Only two protective features were incorporated in older, low capacity systems. Both were located in the cutout and controlled by the operation of a polarized relay. Polarization of the relay prevented the main contactor from connecting the generator to the bus or battery if its polarity built up incorrectly, and a series coil in the same relay caused the relay to open the main contactor in the event of excessive reverse current, preventing battery discharge.

This system and its two protective features worked reasonably well on low capacity and low voltages. On much higher capacities and higher voltages the results are not as favorable. Generator, contactor and cable faults, and overvoltage resulting from faults have caused complete loss of the electric power supply system, and have damaged batteries and other devices sensitive to power system faults as well as involving fire risks.

Several types of faults are possible and all of them have occurred at one time or another. Power systems having simple protective apparatus against these faults have been developed and are coming into use. In addition to protection against battery drain and reversed generator polarity, protection now can be provided against overvoltage from faults or other causes, grounded generator leads, grounded generators, overheated generators, and low voltage resulting from improper equalizer operation caused by wide generator speed differentials and frozen contactors. Trip-free reset may be provided for fault indications. Although these fault devices may operate infrequently, in case of trouble they are of vital importance.

Details of Protective System

Overvoltage

Overvoltage from one or more generators is caused by excessive current in the shunt field windings for the speed and load conditions under which the generator is operating. In the past, the ratio of generator capacity to the battery capacity was small in comparison to the ratio of the larger generator capacity to the battery

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follow rigid rules in all respects. The general aim is to meet the requirements of the particular situations that arise, most effectively. Frequently, perhaps usually, a given end may be attained with equal effectiveness despite considerable variations in the details of accomplishment, so that personal preferences may play some part in the final solution. On the other hand, economic considerations may dictate radically different solutions to apparently similar technical problems at different places. The attempt has been made in this paper to emphasize the rather general aspects of protection; no attempt has been made to analyze technical details exhaustively. It is hoped that we have succeeded in preventing the trees from causing us to lose sight of the forest.

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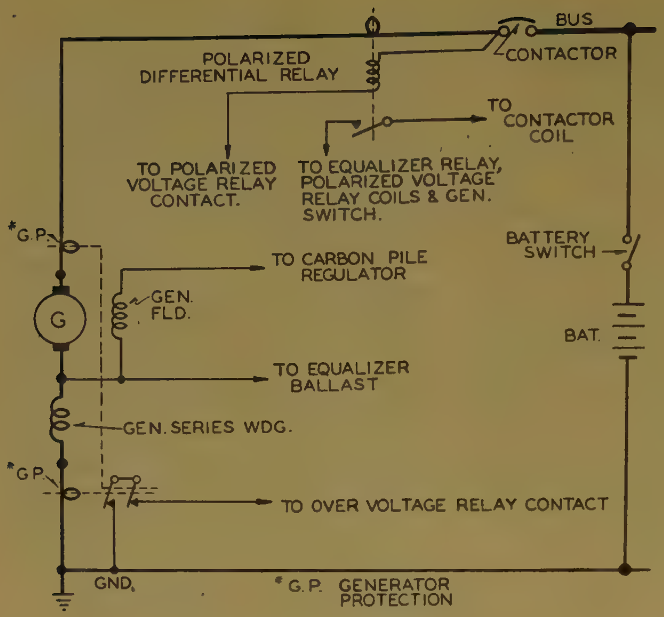


Figure 2. Diagram showing generator and generator lead fault protection

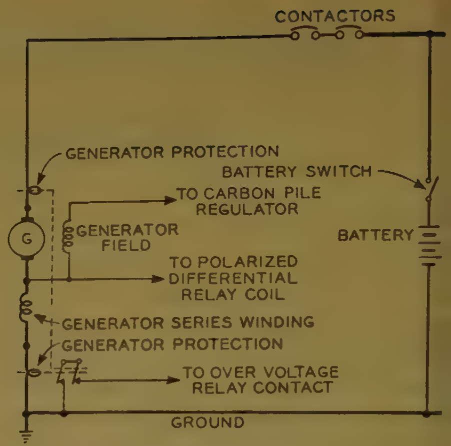


Figure 3. Diagram showing double contactors, generator, and generator lead fault protection

capacity. As a result, the battery of the smaller system could absorb excessive charging current for a much longer time without difficulty and could prevent much increase in voltage. The fault causing the trouble may be sought out and corrected at the next landing. High capacity systems do not have this advantage, and therefore overvoltage protection in larger planes and power systems is certainly desirable and is probably necessary.

Several types of faults are responsible for overvoltage conditions. Wear of carbon pile regulators has been pointed out as an offender. This condition can happen with the older models of this regulator. Design improvements reduced or eliminated this type of fault in modern regulators. However, broken leads, short-circuited terminals, short-circuited leads, or any external fault which will increase the generator field current beyond the control of the regulator will bring about overvoltage.

Relays designed to detect overvoltage operate effectively to remove from the system any generator which is producing overvoltage on the system. Operation of the relay also disconnects the generator field acting through the control devices provided for the purpose. The relay is provided with inverse timing characteristics to prevent transient overvoltage from causing nuisance operation. At 30 to 32 volts its tripping time is one to three seconds, at 60 volts its tripping time is approximately 0.025 of a second. In other words, the higher the overvoltage, the faster the action of the relay. Practically instantaneous operation at extremely high overvoltages is important to avoid damage to electronic equipment such as radio, radar, and similar devices.

After spending considerable time on the investigation of time element relays for overvoltage sensing, a relay with the following description was found which solves the problem in a very reliable manner. It consists of a magnetic piston enclosed within a hermetically sealed nonmagnetic cylinder. The relay coil is wound near the end of the cylinder close to the armature. The piston is hollow, is enclosed at one end, and is normally located toward the end of the cylinder away from the armature. In order for the piston to move, air must flow around the snuggly fitted piston, producing a time lag. When voltage is applied, the piston moves toward the armature, decreasing the air gap. When the piston moves close enough, the armature closes, operating the relay contacts. The higher the applied voltage, the faster the relay operates. A spring returns the piston to its original position when the relay is de-energized.

Multiengine airplanes with one or more generators producing overvoltage will cause the bus voltage to rise, resulting in reverse current to the generators regulated for normal voltage. This condition will cause the main contactors of these generators to trip by the action of the polarized reverse current relay. The voltage of the offending generator will rise still more, resulting in the immediate isolation of the generator from the system. The generators taken off of the bus or battery by the reverse current relay will return to normal operation through the action of the differential relay. Removing the offending generator from the system may clear the fault. The normal operating controls are reset by opening and closing the generator switch. When this is done, the generator will be automatically returned to the system, but if its voltage is still high, the overvoltage relay again

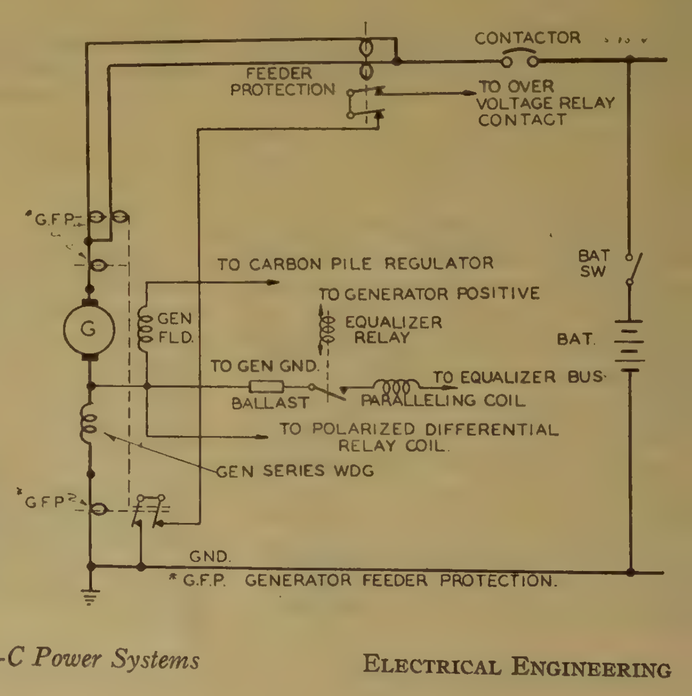


Figure 4. Diagram showing generator, generator lead fault protection plus generator feeder fault protection

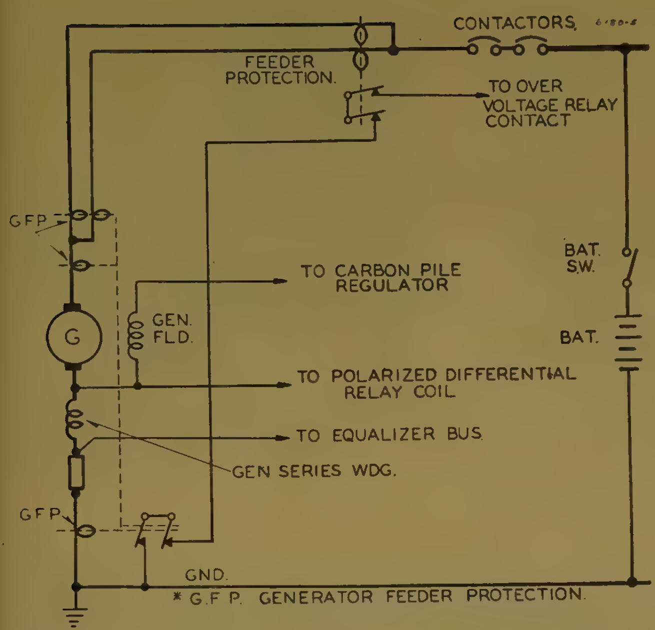


Figure 5. Diagram showing double contactors added to diagram of Figure 4

will remove it from the system. In this case, the flight engineer can decide whether or not he should reset the trip-free control again by opening and closing the generator switch. Two or three trials are considered reasonable.

Generator and Generator Lead Faults

Generator lead faults are infrequent, but probably occur more often than faults within the generator. The generator and generator lead fault protection relay will take care of both types of faults. This relay is usually mounted in the nacelle near the generator. It consists of a single contact operated by two differential series windings on the same magnetic circuit. One of the windings is in series with the positive generator lead, and the second winding is in series with the negative lead. A fault in the generator or at the generator leads unbalances the magnetic circuit of the relay, resulting in the opening of the contacts. This in turn, through the control devices provided for the purpose, isolates the generator and disconnects its field. The trip-free controls may be reset by opening and closing the generator switch. Resetting may be repeated at the judgment of the operator.

Generator Feeder Protection

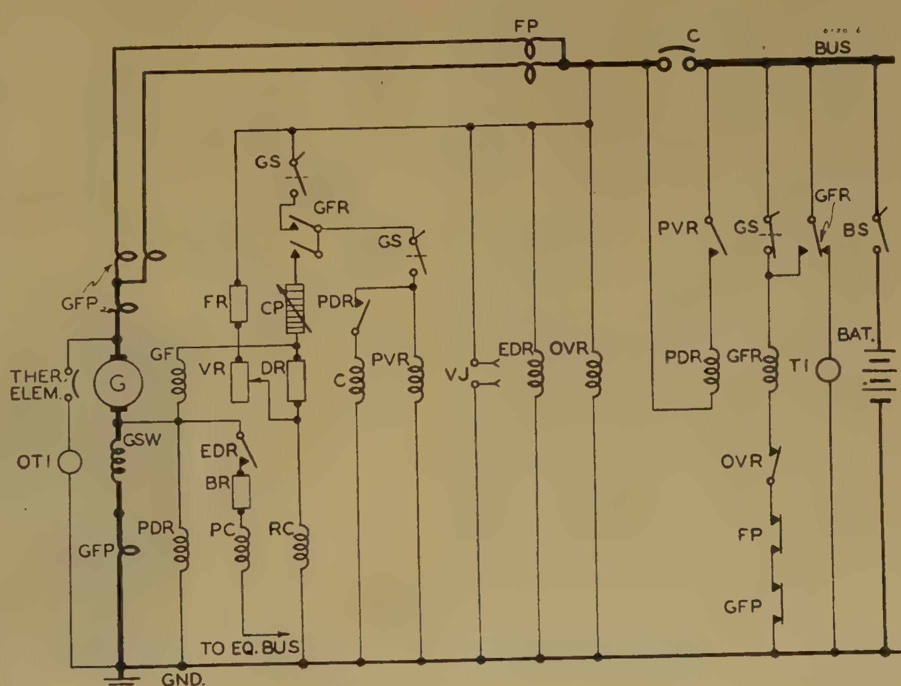
Faults can occur in cable extending from the generator in the nacelle to the bus in the fuselage. Although such faults are very infrequent in well-installed and maintained generator feeder cables, they are extremely serious when they occur. A two-lead differential method is provided for this protection. Two relays are used, each of the differential-series-coil balanced magnetic circuit type. One is located at the nacelle and the second one is located in the fuselage. Each of the two leads passes through one of the series coils of each relay. A fault of predetermined magnitude in either of the two cables will unbalance the magnetic circuit in one or both of the relays and cause its contacts to open. This will cause the control devices to isolate the fault and, at the same time, will disconnect the field of the generator supplying current to the fault. The trip-free controls can be reset at the option of the operator by opening and closing the generator switch.

Current Balance Versus Reverse Current Circuit Breaker

Some degree of generator feeder protection may be obtained by using a reverse current circuit breaker adjacent to the main contactor. When a major generator feeder fault occurs while the main contactor is closed, the circuit breaker will trip on a very heavy reverse current entering the feeder from the bus or the battery. This trip setting is normally about 50 per cent greater than the rating of the generator with a time delay, and instantaneous at currents 100 per cent above the trip setting. The high minimum trip current and the time delay feature are necessary to prevent false tripping during momentary reverse current conditions imposed by system voltage errors or by generator re-

Figure 6. Schematic diagram showing the complete control and protection of a single aircraft d-c generator

BR	Ballast resistor
BS	Battery switch
C	Contactor
CP	Carbon pile regulator
DR	Damping resistor
EDR	Equalizer disconnecting relay
FP	Feeder protection
FR	Fixed resistor
G	Generator
GF	Generator field
GFP	Generator feeder protection
GFR	Generator field relay
GS	Generator switch
GSW	Generator series winding
OTI	Overtemperature indicator
OVR	Overvoltage relay
PC	Paralleling coil
PDR	Polarized differential relay
PVR	Polarized voltage relay
RC	Regulator coil
TI	Trip indicator
VJ	Voltmeter jack
VR	Variable resistor



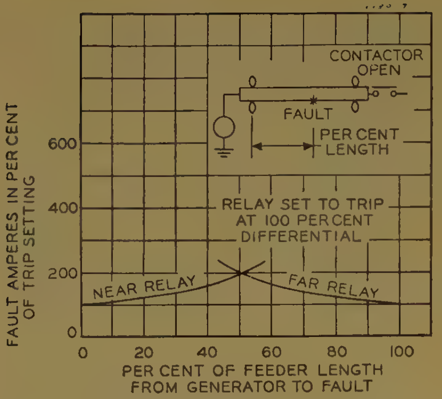


Figure 7. Curves showing fault amperes in per cent of trip setting of the protective relays for a typical generator feeder

weight than current balance and double contactors having remote reset control.

Double Break in Generator Bus Feeder

A single main contactor for connecting the generator to the bus or battery has been used extensively. There are important reasons for using two identical contactors in series with their operating coils in parallel. Because of slight manufacturing differences in magnetic circuits, springs, and mechanical parts they never may open and close at the same instant. One of them normally will operate, leaving the other as a backup unit

The backup contactor will be available for instantaneous operation if the other unit fails to open. Frozen contacts may not cause trouble during flight, but after landing and shutdown of engines the pilot may fail to open the battery switch. In such a case the frozen single contactor will cause complete discharge of the battery. This might not damage the generator; however, if an airport power supply is plugged in, the generator is certain to burn out.

Contactors having new features which reduce the chances of freezing are in use. They have high pressure at the contacts, inertia backup during closing, and a very low voltage drop-out. With such contactors, danger of freezing of contacts is slight and, for all practical purposes, the use of double contactors will eliminate these risks to the electric power system.

Ballast Equalizer Connections

A system with two or more generators requires provisions for satisfactory parallel operation. The only methods now in use consists of coil windings in each regulator, operated from the drop across an external resistor in series with the generator or indirectly from a series winding inside the generator. The series resistor usually provides one-half of a volt drop at full load current and has been used extensively for military aircraft. Paralleling has been moderately satisfactory. However, paralleling for commercial transports should be good enough to prevent the overloading of one or more generators in flight or ground operations. Stand-by loads for 2-engine airplanes require reasonably accurate paralleling because this operating condition is likely to heat the generators more than any other phase of operation as a result of limited generator cooling air. Four-engine airplanes require paralleling to be reasonably accurate for all operating conditions because three generators must run at full output

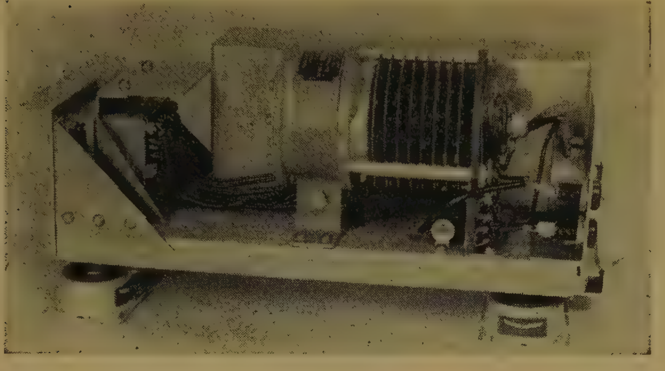


Figure 8. Control panel including regulator plugged into shock mount

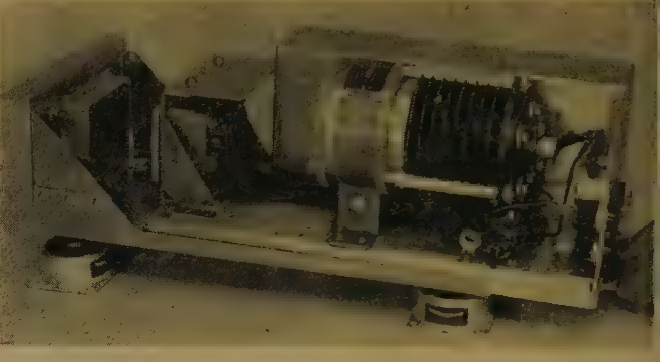


Figure 9. Control panel shown in Figure 8 being removed as a single unit from the shock mount

Note separation of plug and receptacle

covery characteristics following an external load circuit fault. As a consequence, it is quite possible to have a generator feeder fault of a magnitude that will open the contactor by reverse current action before the circuit breaker can trip. If this occurs, the generator field will not be opened and the generator will continue to feed the fault. A similar condition will exist if a fault occurs while the contactor is open, or if the generator builds up on a faulted generator feeder.

The well-known current balance feeder protection is not subject to these errors. Because it is not responsive to normal load current, reverse current, or external fault conditions, it can be set to trip on faults considerably lower in magnitude than the current setting limitation of the reverse current circuit breaker. Most important, current balance protection operates without dependence on reverse current relay sequence, timing devices, or main contactor status. If the contactor is already open, the protective relay proceeds to open the generator field, and thus prevents feeding the fault. Manual reverse current circuit breakers are slightly less in

Figure 10. Control panel shock mount showing clamp and receptacle

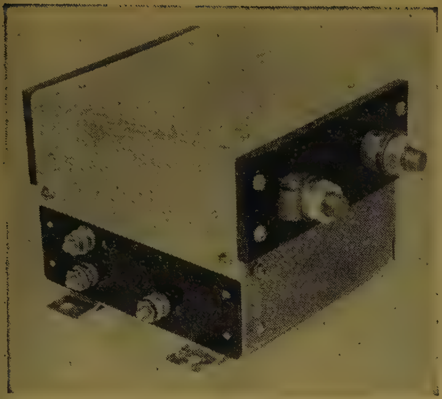


Figure 11. Nacelle type contactor and ground protection relay box

in case of the failure of any one generator.

Commercial transports are subject to delayed ground operations at many airports. The equal loading of each generator under these conditions is important. Increased equalizer voltage increases the sensitivity of the regulator to unbalanced current conditions and provides much better parallel operation. Full voltage is being taken from across the series winding of the generator and is approximately $1^1/_2$ volts.

Increased equalizer voltage also results in a lower system voltage when the speed of one or more generators is lowered to the idling value. This is an unfavorable condition, and it is corrected to some extent by a nonlinear ballast in the equalizer circuit, which limits the equalizer current under wide generator speed differences.

An equalizer disconnecting relay is also provided which opens the equalizer connection between regulators when fault protective devices operate or when the generator switch is opened. This then permits all other generators to operate at normal voltage when any one or more generators are cut out as a result of faults or shutdown. The relay is a simple voltage operated 2-position type.

Generator Thermal Indication

Overheating of the generators may be caused by overloading or insufficient ventilation at normal loads. Insufficient ventilation usually occurs under stand-by conditions. Analysis has shown that the only satisfactory device for indicating overheating of the generators is a thermal indicator sensitive to temperature and independent of current or load. The generator is not disconnected from the system when overheating is indicated, as modern insulation permits the generator to operate until the nonessential loads can be removed and the cause investigated at the next landing.

Stabilized Voltage Regulators

This type of regulator consists of a stack of carbon and graphite disks contained in a heat-radiating cylinder and a solenoid, or regulating coil, matched over a wide range to a spring which is arranged to apply varying pressure on the carbon pile in accordance with the voltage in the generator circuit. The carbon pile is connected in series with the generator field circuit so that the resistance variation through the carbon pile causes variations in the generator field current, thus obtaining constant generator voltage regardless of connected load or speed. The stability of the regulator is achieved by a damper mechanism employing both air and spring damping. Tests have demonstrated stable operation throughout the life of the regulator.

Control Apparatus Assembly And Locations

For elementary systems used on military aircraft and some commercial transports, the regulator is usually mounted in the fuselage, and the differential relay and ground resistor mounted in the engine nacelle. This is a reasonably satisfactory arrangement for this type of apparatus.

Systems provided with protective devices as described in this paper are designed with considerable flexibility as to the assembly of each unit and the mounting locations. Regulators and relays are preferably assembled on a plug-in base similar to the plug arrangements used with radio panels and racks. With this arrangement the major control units can be replaced in a few seconds. The assembly of the fault protection relays and the contactors are generally in single units and are located to the best advantage in the particular airplane being constructed.

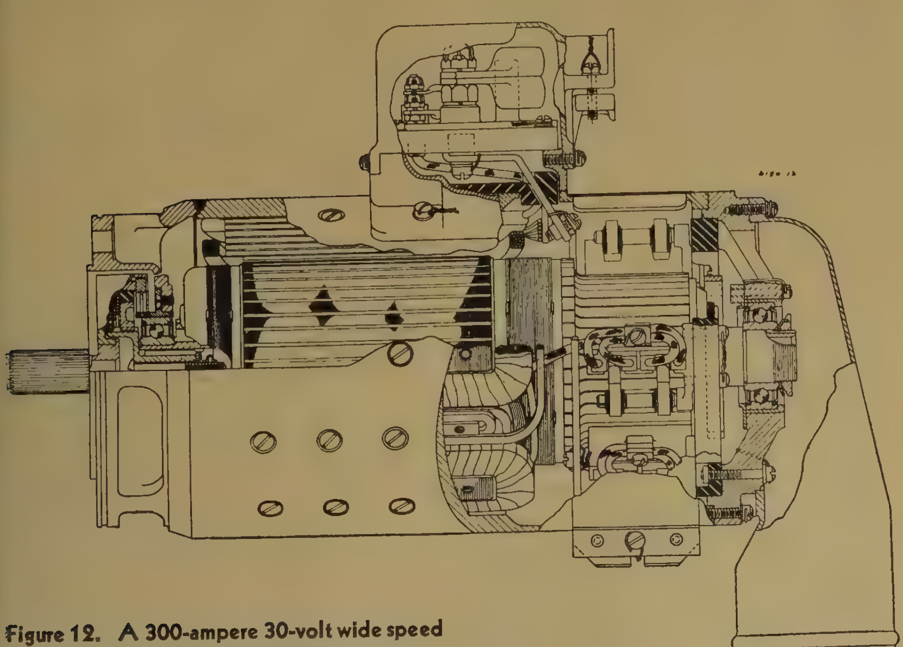


Figure 12. A 300-ampere 30-volt wide speed generator showing internal arrangements

Conclusions

Many of the simple principles used in the aircraft d-c power and protective systems are those used extensively in land power systems where continuity of service is of first degree importance.

Cable, generator, overvoltage, frozen contactor, and generator overheating faults are isolated or indicated during all operating conditions. Otherwise the system functions normally and fully automatically to connect and disconnect the generators from the battery bus as required. Satisfactory operation under conditions of humidity, sand, dust, vibration, altitude, explosive gas fumes, and acceleration normal to aircraft has been achieved by designing with these extreme conditions in mind.

Means for quick removal and replacement of the complete control panel by using standard radio plugs and receptacles provides the ultimate ease in maintenance and involves no compromises with performance. Contactors for any given power system are designed and applied so that they will interrupt successfully without undue stress any top current value that can be obtained on the system at the point where the contactor is located.

Reference

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Protection of Powerhouse Auxiliaries

AN AIEE COMMITTEE REPORT

THE AIEE relay subcommittee has reviewed the protection of generating stationauxiliary power motors and supply systems. The study was made to obtain information on auxiliary power systems now in use and the protective methods being applied, and to make recommendations on preferred practices.

In order to obtain basic information, a questionnaire was sent to approximately 30 engineers interested in this subject. These questionnaires covered data on the supply systems used, various forms of protection applied, and a summary of the experience obtained with these systems. Answers to the questionnaire were received from various representative companies covering both steam and hydro stations and a wide range in station sizes. The answers submitted represented the modern practice of each company. This subject also was discussed in a conference session held during the 1946 winter convention.

Summary of Information Received

Following is a brief summary of the information received on the various points covered by the questionnaire.

Supply Systems Used

In most cases, transformers are being used for auxiliary power supply. Values of 2,300 and 440 are the principal voltages used, with motors higher than approximately 100 horsepower being supplied by the higher voltage system. In most cases the systems operate ungrounded, although some companies prefer to operate grounded, either solidly or through impedance.

Paper 46-169, recommended by the AIEE committee on protective devices for presentation at the AIEE summer convention, Detroit, Mich., June 24–28, 1946. Manuscript submitted April 19, 1946; made available for printing May 29, 1946.

This report was prepared by a working group on powerhouse auxiliaries of the relay subcommittee, AIEE committee on protective devices, consisting of E. L. Michelson (M '44, sponsor), senior engineer, Commonwealth Edison Company, Chicago, Ill.; A. H. Burkhalter (M '41), electrical engineer, design and construction department, Ebasco Services, Inc., New York, N. Y.; R. E. Cordray (M '43), in charge of relay engineering, General Electric Company, Philadelphia, Pa.; E. H. Klemmer (A '42), switchgear engineer, Westinghouse Electric Corporation, East Pittsburgh, Pa.; W. E. Marter (A '40), assistant relay protection engineer, Duquesne Light Company, Pittsburgh, Pa.; C. L. Smith (A '28), relay engineer, electric distribution department, Rochester Gas and Electric Corporation, Rochester, N. Y.

The protection of the auxiliary power transformers includes, in most cases, differential relays with time delay overcurrent protection for backup purposes. In many installations there are two sources of primary supply with an automatic throwover arrangement, initiated by undervoltage relays, which connects the reserve supply if the primary source is lost.

On busses supplied by auxiliary generators, the problem of bus protection is complicated by the decrement in fault current supplied by the generator. The decrement causes a rapid reduction of fault current to a point approaching normal load current, so that it is difficult to obtain bus protection with time delay overcurrent relays. In such cases differential or fault bus relays provide the most desirable method for bus protection.

Motor Protection

In the majority of cases, the 2,300-volt motors are protected by long-time-delay overload relays set at 125 to 150 per cent of full load current and instantaneous overcurrent relays on the cable leads which are set above the starting inrush current. The smaller motors are in general protected by thermal tripping elements for overload, and an instantaneous trip device for short circuits.

There is some variation in the treatment of essential and nonessential motors. In most cases, the essential and nonessential motors are protected by similar relays. However, in a few cases the circuit breakers feeding essential motors are not tripped by the time delay overload relays. These relays give an alarm only in the event of an overload condition.

Essential motors are defined as those motors whose failure results in the shutdown of generating capacity. Examples of this class of motors include the circulating pump, boiler feed pump, and induced and forced draft fans.

Other Forms of Protection

In most cases where the systems are operated ungrounded, ground detectors are available for detecting grounds.

In powdered fuel plants, it is necessary to interlock the sequence of starting and stopping certain motors. For example, the induced draft fan is started first, and is followed in sequence by the forced draft fan, the exhauster, the pulverizer, and the feeder. If any piece of equipment in this series is stopped, provision is made to stop automatically all of the motors ahead of it.

Experience and General Conclusions

The answers received indicated that as a whole the protective schemes now in use are satisfactory. The only indication of unsatisfactory performance has been with the thermal relays used on the smaller motors. Various types of troubles have been reported, such as lack of thermal capacity to withstand fault currents, and lack of sensitivity. One company uses oversize thermal elements to avoid difficulties which have been experienced with the sizes recommended by manufacturers.

In general, the importance of auxiliary power in generating stations is so fundamental that simple and rugged relaying schemes are to be desired. In almost every case the auxiliary power supply is the last element of the system which should be shut down, and therefore the relaying systems used should be as free from trouble as possible.

Recommended Preferred Practices

Supply Systems

For ungrounded systems or grounded systems where the ground fault current is limited to a few amperes, ground detector relays should be used.

For systems grounded solidly or through low impedance, residual current relays should be used on individual circuits.

Transformers supplying essential auxiliary power busses should be protected with differential relays, and time delay overcurrent relays should be used for backup protection.

Bus protection should be provided by overcurrent, differential, or fault bus relays. Overcurrent relays on auxiliary generators should not be relied upon for bus protection.

Motor Protection

The 2,300-volt motors should use long-time-delay phase overcurrent relays for overload and internal motor faults set at approximately 150 per cent of rated current. They also should be equipped with instantaneous overcurrent relays for short-circuit protection set above maximum inrush current.

On essential motors the time-delay overload relays may be used for alarm purposes only, and the instantaneous relays used to trip. In this case, the time delay relays can be set more sensitively than 150 per cent of rated current.

Low voltage motors (208, 440, 550 volts) should use thermal device for overload protection, and an instantaneous trip device for short-circuit protection.

The thermal device may be built into the contactor supplying the individual motor, and one circuit breaker may provide short-circuit protection for a group of motors.

On essential motors the thermal element may be used for alarm purposes only.

Answers to Questionnaires

Following in a summarized form are the individual answers received.

Company Number 1

Supply System Used. This company uses transformers for supplying power to

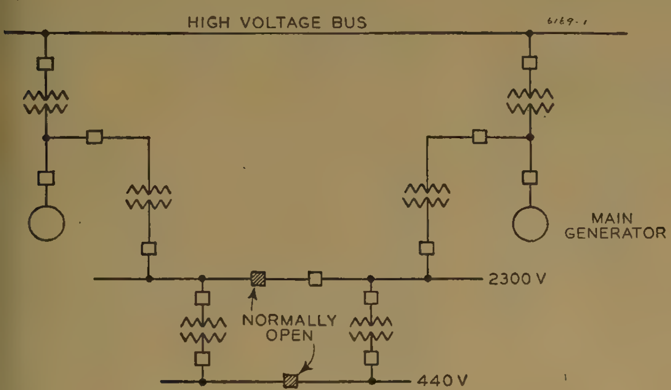


Figure 1. The supply system used by company number 2

Differential and overcurrent protection on auxiliary transformer. Differential closes bus tie when transformer breaker trips

2,300- and 440-volt busses. The transformers are operated ungrounded.

The transformers are protected by overall differential relays with time delay overcurrent protection on the low voltage output and time delay phase and instantaneous ground protection on the primary.

Motor Protection. Motors of 75 horsepower and higher use 2,300 volts. These motors have long-time-delay induction overload relays set at about 150 per cent of full load. Instantaneous relays also are used and are set to exceed slightly the maximum starting inrush current.

The 440-volt motors all are protected by circuit breakers with thermal tripping elements.

Other Forms of Protection. Static type ground detectors or lamps are used to detect grounded circuits, which are located by opening the circuits in turn.

Most essential auxiliaries are equipped with dual drive.

Both 2,300- and 440-volt busses are arranged for automatic throwover to adjacent sections in case of loss of potential, although the throwover will not function in case of a fault on the bus. Throwover is initiated when the voltage drops to about 65 per cent.

Experience. As a whole the protection has been satisfactory.

General Conclusions. It is felt that the ideal type of auxiliary power motor protection would consist of an instantaneous relay sensitive to all motor faults but insensitive to overloads, plus an overload relay with good correlation between its operating characteristic and motor temperatures.

Company Number 2

Supply System Used. The supply system used is illustrated in Figure 1.

Motor Protection. Motors 75 to 100 horsepower are supplied at 2,300 volts. On 2,300-volt essential motors, the protection consists of instantaneous overcurrent plunger relays for short-circuit protection. Thermal relays give visual indication only. On nonessential motors, the protection consists of 40-second overcurrent induction relays with instantaneous attachments.

On 440-volt essential motors, the protection consists of small instantaneous air circuit breakers. Thermal elements of motor contactors only light lamps. The nonessential 440-volt motors have the same protection except that the thermal elements trip the motors.

Other Forms of Protection. Ground detector relays are used on the 2,300-volt system and lamps on the 440-volt system.

An interlocking system between motors is used on the boiler auxiliaries to protect the boiler against motor failures.

Experience. The reliability of modern motors justifies a minimum of protection and maximum freedom from undesired tripping.

Relays for 2,300-volt breakers are satisfactory, but contactor elements, small air circuit breakers, and so forth, for 440-volt service are not sufficiently reliable.

Many applications require "latched-in" contactors to prevent disconnection on voltage surges.

Company Number 3

Supply System Used. The supply system used is illustrated in Figure 2.

Motor Protection. Motors up to 125 horsepower are connected to the 440-volt bus.

All 2,300-volt motors are protected by

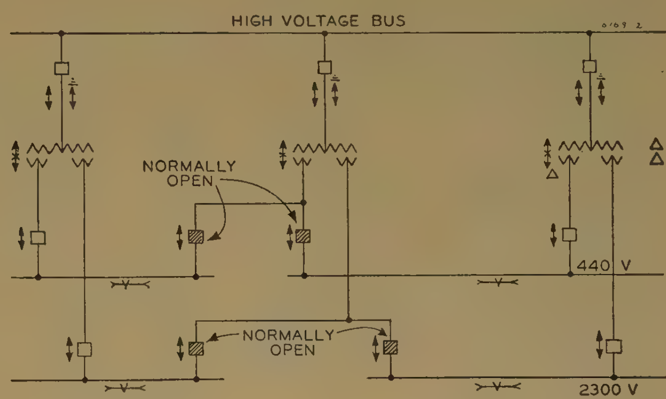


Figure 2. The supply system used by company number 3

Undervoltage relays throwover to reserve transformer

differential relays using instantaneous overcurrent relays set for approximately 50 per cent of full load current.

On the 440-volt system, feeders from the main bus are used to supply subbusses throughout the plant. These feeders use circuit breakers with a series trip set for 200 to 400 per cent of the rating of the breaker. These breakers protect against faults on the subbusses and heavy faults on the feeders to the 440-volt motors. Each motor is protected individually by a contactor with a thermal relay set for 30 seconds at 150 per cent load.

Other Forms of Protection. Lamps are used for ground detecting system. Grounds are removed by disconnecting portions of the system until the ground is located. The location of grounds is delayed until it is convenient to take equipment out of service.

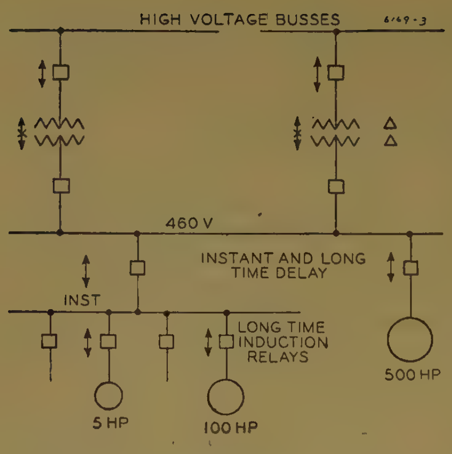


Figure 3. The supply system used by company number 5

Experience. High speed clearing of faults has resulted in very satisfactory operating conditions and has proved advantageous in limiting the extent of damage caused by motor faults.

Lack of thermal overcurrent protection on 2,300-volt motors has resulted in some difficulty with detecting cases of overload on mill motors.

Operation of thermal elements on 440-volt motors has been satisfactory.

General Conclusions. While differential protection on 2,300-volt motors has proved to be advantageous, it is questionable whether their use is economically justifiable. Future installations will use overcurrent relays.

Company Number 4

Supply System Used. The supply to auxiliary power is from transformers 13.8/2.3 kv, delta-Y connected. The 2.3-kv neutral is grounded solidly. Smaller motors are supplied from a 575-volt system, which in some cases is grounded and in other cases ungrounded. The transformers are protected by differential and induction type overcurrent relays.

Motor Protection. Motors up to 100 horsepower are supplied from the 575-volt system.

Motors are protected by a 40-second induction-type relay with instantaneous attachment. The time delay element is set at approximately half way between the ordinates of the motor full load temperature curve and the curve representing motor thermal limitations. The instantaneous trip is set just over "locked rotor" current. In some instances, thermal relays are used which are set in the same manner as for the induction relays.

No distinction is made between methods employed for protection of essential and nonessential motors except that the instantaneous auxiliary overcurrent relays set at approximately 115 per cent of the rated current are applied to pulverizer motors. These relays stop the coal feeder motors upon overload of the pulverizer until the pulverizer motor current drops back to normal.

Other Forms of Protection. Very elaborate interlock systems are used between induced and forced draft fans, pulverizers, boiler feed water pumps, and so forth.

Ground lamps are used for ground protection. No grounds have developed to date.

Experience. Performances of protective and control schemes have been excellent. All thermal relays of a certain make have had to be dismantled and reassembled in our laboratory before being placed in service because of poor factory workmanship.

Speeds of clearing faults have been very satisfactory; all cable and motor fault currents have been of a magnitude high enough to operate the 1-cycle relays. Iron and winding damage has been negligible.

General Conclusions. The simplicity of the 40-second induction overcurrent relays with instantaneous current trip features makes this scheme a slight favorite over the thermal combination for 2.3-kv motors. The manufacturer's standard protection on 575-volt cubicle switchgear has been found satisfactory.

Company Number 5

Supply System Used. The supply system used is illustrated in Figure 3.

Motor Protection. All motors are supplied at 460 volts (5 to 500 horsepower).

The 40-second induction overcurrent relays are used on almost all auxiliary motors. Current settings are from 125 to 150 per cent of motor rating. Time settings are generally about one second minimum time.

Instantaneous trips are used on all group breakers to clear any short circuit of sufficient magnitude to affect the operation of other motors.

Other Forms of Protection. Mill motors and exhauster fans are interlocked. Coal conveyer systems are all interlocked.

A set of ground indicating lamps with series resistors is installed on each auxiliary bus. Ground is established, for test, by a push button. This is a routine test performed by the operator on each shift.

A ground is a rare occurrence on the 440-volt busses, but as soon as one do occur it is usually possible to cut in dup cate equipment and clear the trouble.

Experience. Our experience wi thermal relays has not been satisfactor The testing and maintenance cost has be extremely high. We have found that t long time induction overcurrent rela though slightly more expensive initiall is far more reliable, maintains its calibr tion, and is easy to test. Quite a few these motors are equipped with therm guard protection. We have not use these devices for tripping, but have co sidered the possibility of doing so.

General Conclusions. Differenti protection is not justified on 440-vo auxiliary motors because winding failur are rare.

Instantaneous tripping should be pr vided for every group or feeder circu breaker which is connected to the mai auxiliary bus.

Company Number 6

Supply System Used. The suppl system used is illustrated in Figure 4.

A 440-volt system is used which is fe from either of the 2,300-volt busses.

Motor Protection. On motors fe from the 2,300-volt service, long tim overcurrent relays with instantaneou elements are used. Long time overcur rent relays having a time delay up to 4 seconds are used and can be set for 11 to 125 per cent of full load current an with sufficient time delay to ride over th starting inrush. The instantaneous ele ments are set above starting inrush an furnish protection for faults on moto leads and circuit.

The main 440-volt motors are equippe with individual circuit breakers with ther momagnetic overload protection. Thi

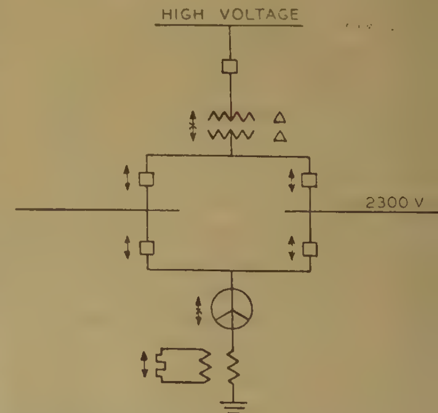


Figure 4. The supply system used by com pany number 6

provides both overload and short-circuit protection for these motors. Time delay undervoltage type *CV* relays are used on the 2,300-volt motors.

Other Forms of Protection. Automatic transfer of busses from the transformer to the generator source is provided to transfer load in case of voltage failure. The transfer scheme is interlocked to prevent paralleling of busses or to transfer after an overload relay operation.

The 440-volt busses have automatic transfer between transformers connected to the two 2,300-volt bus sections and have interlocks similar to the 2,300-volt busses.

Experience. The protective schemes used have given very satisfactory service

Table I

Motor Voltage	Max Hp	Min Hp	Protected by
2,300	1,000	300	Individual circuit breakers
550	250	50	Individual circuit breakers
550	50	1/2	Magnetic contactors

and no load curtailment or shut down has occurred because of station service failures. The equipment described is new and few failures have occurred.

General Conclusions. The long time overcurrent relays with instantaneous attachments seem to offer satisfactory protection for the 2,300-volt motors since they can be set low enough to prevent roasting of insulation on overload, and the instantaneous elements provide quick clearing for severe faults.

Company Number 7

Supply System Used. The auxiliary power is supplied by transformers at 2,300 and 550 volts. The supply busses are operated ungrounded.

Protection consists of differential relays on the transformer banks and time delay induction overcurrent relays on all supply transformer circuit breakers.

Motor Protection. Motors are supplied as indicated in Table I.

Motors controlled by individual circuit breakers have

1. Instantaneous overcurrent relays set slightly above motor starting current.

(*a*). On 2,300-volt motors these are plunger type relays actuated from current transformers.

(*b*). On 550-volt motors these are built into breaker housing.

2. Time delay overcurrent relays set to pick up at 125 per cent of motor rated current and with sufficient time delay so as not to trip during motor starting.

(*a*). On 2,300-volt motors these are induction type operated from current transformers.

(*b*). On 550-volt motors these are thermal devices built into the breaker.

3. Thermal devices built into the motor structure which open a contact when the motor temperature exceeds a safe value and sound an alarm.

Motors controlled by magnetic contactors have

1. Air circuit breaker for instantaneous short-circuit protection.

2. Thermal overload built into the magnetic contactor.

All motors controlled by circuit breakers are treated as the foregoing.

1. Essential motors controlled by magnetic switches have thermal overload two sizes above that recommended for the size of motor involved.

2. Nonessential motors controlled by magnetic switches have thermal overload one size above that recommended for the size of motor involved.

Other Forms of Protection. An alarm is provided on each ungrounded bus to inform operator when a ground occurs. It consists of double-Y potential transformers with the neutral grounded on the high voltage side and overvoltage plunger type instantaneous relays connected from phase-to-neutral of the secondary winding. When a ground occurs the two relays on the ungrounded phases pick up and sound an alarm.

Experience. Fortunately the occurrences of short circuits on the auxiliary power system have been few and far between. In cases where short circuits have occurred, the instantaneous overcurrent relays in general have operated to clear the circuit with very little resulting damage to the equipment. Also, for *overload* conditions few cases have been experienced, for in general the size of motors chosen have had ample capacity to take care of the maximum mechanical load demanded. In the few cases where overload conditions occurred, the time delay induction overcurrent relays have done a good job without serious damage to the motors. In case of the thermal devices the experience has not been quite as good but not serious enough to change the scheme of protection.

General Conclusions. The use of ungrounded auxiliary power system employing a ground fault detecting system and eliminating the ground immediately after it occurs reduces to a minimum the number of short circuits that occur on the system. Further, the use of instantaneous overcurrent relays do a very satisfactory job when such short circuits do occur. In the use of instantaneous overcurrent relays, care should be taken that they are set at their minimum pickup value (that is just a little above the starting current) and not the general value sometimes recommended of ten times the circuit breaker rating. Also, the employment of induction overcurrent relays set at 125 per cent of motor rated current appears to be more reliable than other forms of overload protection.

Company Number 8

Supply System Used. The supply system used is illustrated in Figure 5.

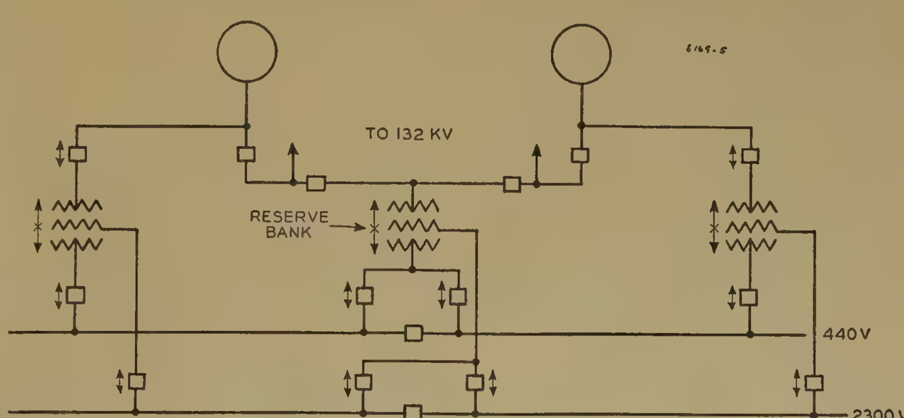
Motor Protection. The 2,300-volt system supplies motors from 175 to 600 horsepower, the 440-volt system 1/2 to 200 horsepower.

Motor protection consists of the following:

1. On all 2.3-kv equipment—one overcurrent long-time relay with instantaneous attachments and two instantaneous overcurrent relays, two current transformers with an overcurrent relay connected in the residual circuit of the two current transformers.

2. On all 440-volt equipment—air circuit breakers (deion) with time delay and instantaneous relays.

Figure 5. The supply system used by company number 8



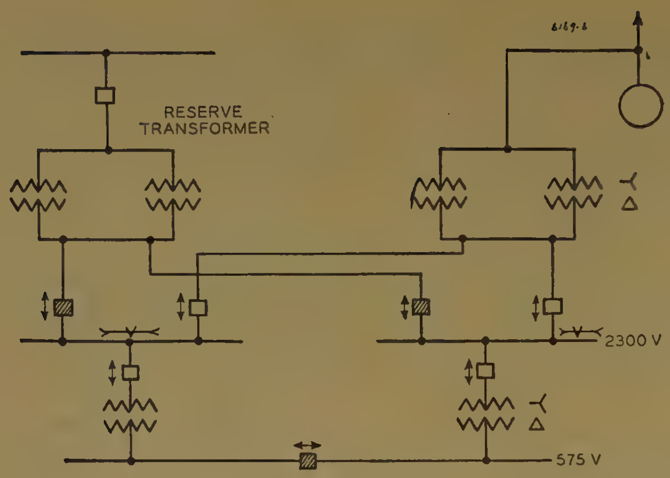


Figure 6. The supply system used by company number 9

Undervoltage relays throwover to reserve transformers

3. Undervoltage type *CV* relays on 440-volt bus trips pulverizer mills.

4. Most 440-volt motors have individual air circuit breakers with thermal trips.

No preference is given between essential and nonessential circuits on relaying procedure.

Automatic throwover of both 2.3-kv and 440-volt busses is provided.

Other Forms of Protection. Interlocks are used on pulverizers and coal scales, also on forced draft and induced draft fans.

Ground detectors are located on both 2.3-kv and 440-volt busses.

Experience. Present relays are very satisfactory. However, those originally installed (two overcurrent standard relays with instantaneous attachments) could not be set to co-ordinate properly with motor starting. Faults have been cleared very quickly and no extensive damage has resulted to motors or cables.

General Conclusions. The afore-mentioned scheme (one overcurrent and two instantaneous overcurrent relays) seems to us to be a very workable and efficient scheme.

Company Number 9

Supply System Used. The supply system used is illustrated in Figure 6.

Motor Protection. The 2,300-volt system supplies motors over 100 horsepower. All 2,300-volt motors have instantaneous overcurrent relays for short-circuit protection. The essential motors have a long-time-delay relay for overload protection which gives an alarm only. On nonessential motors the long-time-delay overload relay trips the motor. On the 575-volt motors, there is an instantaneous trip for short-circuit protection and a thermal trip for overload.

The instantaneous relays are set approximately ten per cent above maximum motor inrush current, that is, ten per cent above asymmetrical peak value occurring in the first half cycle.

Other Forms of Protection. Boiler fan motors and mill motors are interlocked for sequence starting and stopping.

Ground relays on 2,300-volt and 575-volt busses operate an annunciator. Grounds are removed by disconnecting one motor feeder at a time until the ground is located.

Experience. Provision has been made to trip almost all of the motors immediately in case of a short circuit in the feeder or the motor. In faults of this nature the amount of damage is limited by the quick tripping. In a recent case of motor failure it was possible to remove the faulted coils from the circuit by using jumpers and keep the motor in service.

Some difficulty is experienced in co-ordinating the instantaneous direct acting trips of the 575-volt circuit breakers. Where these breakers are cascaded a short circuit usually trips all of them in the cascade with loss of unfaulted motors or feeders. Dashpot devices have not proved satisfactory for this purpose. A simple device, such as an adjustable escapement, on the instantaneous trips for a few cycles timing would be of material benefit in solving the problem.

General Conclusions. In general, th[e] protective scheme adapted is considere[d] reasonably economical and simple and ha[s] performed satisfactorily with the excep[-]tion of the cascading referred to in th[e] foregoing paragraph.

Company Number 10 (Hydroeletric)

Supply System Used. The suppl[y] system used is illustrated in Figure 7.

Motor Protection. All motors ar[e] supplied at 550 volts. The motors ar[e] supplied through individual "Non-fuz[e]" breakers. Between each of the breaker[s] and the individual motors, there ar[e] magnetic switch starters with a suitabl[e] thermal relay for overload protection.

Other Forms of Protection. The sta[-]tion service is not grounded and there is n[o] means of detecting and removing ground[s].

Experience. No difficulty has bee[n] experienced with any part of the statio[n] service.

Company Number 11 (Hydroeletric)

Supply System Used. All auxiliar[y] power motors are supplied at 550 volt[s]. The supply is from transformers tha[t] are fed from house generators. Thes[e] transformers are protected by overcurren[t] relays. The auxiliary power suppl[y] system is grounded solidly.

Motor Protection. All auxiliary mo[-]tor circuits are protected by overcurren[t] inverse time relays, plus overcurrent i[n]stantaneous elements at a higher settin[g].

Other Forms of Protection. No othe[r] forms of protection are used.

Experience. Experience in operatin[g] for 13 years with this protection syste[m] has not shown any defects in the metho[d]. However, no essential auxiliary motors [or] circuits have developed a fault thus fa[r]. The crane motor-generator set and tw[o] sump pumps are the only equipment to d[e]velop any trouble, and the trouble was n[ot] extensive. A coil or two short-circuite[d]

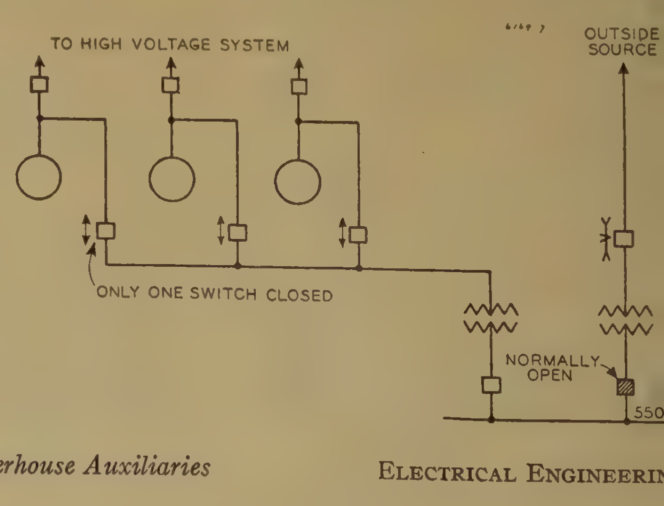


Figure 7. The supply system used by company number 10

Six Years' Experience With Factory-Built Unit-Type Substations

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THE Iowa-Illinois Gas and Electric Company installed its first unit-type substations late in 1939. At the present time, there are installed in the various operating districts of the company nine unit-type substations. We are outlining the reasons for the choice of this type of substation and the distribution system into which the substations are being fitted. This paper will cover our operating experience with six of this type of substation located in the quad-city area of Davenport, Iowa; Rock Island, Moline, and East Moline, Ill. The reasons given for the selection of this type of substation, the design of the distribution system served by the substation, and operating experience so far encountered are typical also of the other districts in which the additional three substations are located.

The quad-city area comprising the cities outlined above occupies two states, Iowa and Illinois. The division between states and cities is made by the Mississippi River. Davenport, Iowa, and its suburban towns, occupying the north bank of the Mississippi River and located in the State of Iowa, comprise a total population of approximately 90,000. The cities of Rock Island, Moline, and East Moline, Ill., with their attendant suburbs located on the south bank of the Mississippi River, comprise a total population of approximately 114,000.

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Because of the advantages offered by the Mississippi River, the bulk of industry is concentrated along the river banks. The cities of course started their growth on the banks of the river and have expanded outward. Greatest load concentrations are along both banks of the river with the residential load being located in an approximate semicircular area around the outside of the industrial area. Original substations are located close to the river in the heavy industrial and commercial areas.

Polyphase alternating current was introduced into the quad-city area in 1888, six years after the first electric service was installed. This early a-c system was 2-phase, and, as a result, all growth on the quad-city system for many years was 2-phase. This 2-phase service was utilized in two ways.

1. The 2-phase 4,800-volt primary circuits were used to supply industrial customers.
2. The 2-phase 2,400-volt circuits were used to supply lighting customers.

The 2,400-volt system was utilized throughout much of the residential area as single phase primaries. The distribution system remained single phase 2,400-volt and 2-phase 4,800-volt until late in 1939. In the meantime, a backbone transmission and subtransmission system of 69-kv and 13.8-kv 3-phase circuits was built which served the complete quad-city area.

The later load growth has tended to be away from the river to keep pace with the expanding residential areas and some smaller outlying industrial areas. Because of growth of the communities, the load center of residential load kept shifting out away from the river and the original manually attended substation locations. This change in load center was followed by lengthening of the distribution lines until the long lines introduced poor regulation despite the use of induction feeder voltage regulators. Interruptions of service on these long feeders involved a large number of customers. As the areas grew, a number of small outlying commercial areas developed in which only single phase service was available. This offered many disadvantages to the merchants located in these areas as much of the equipment which they purchased normally came equipped with 3-phase motors.

The requests for polyphase service were each of relatively small electrical demand and scattered over fairly large areas. As a result, it was uneconomical to construct long polyphase feeders from the existing substation locations to serve the widely diversified loads of small magnitudes. The many requests for polyphase service received from the outlying areas required that consideration be given to the establishment of polyphase primaries in these localities. The location of the existing substations and number of circuits already leaving those stations made it difficult to get out additional circuits without going considerable distances underground, thereby raising the cost of each of the outgoing circuits which would be required.

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in the sump pumps which caused tripping by the overload protection during running, and clearing took place quickly enough to prevent any spreading of the fault. The crane motor-generator set became defective because of broken coil leads caused by insufficient bracing to withstand line start operation. The overload relays occasionally would trip out during the starting period for some time before the condition developed in enough bars to prevent satisfactory operation. No burning or spreading of the damage resulted, and it was difficult to locate the flaws which had occurred in the winding.

It should be remembered that in line start operation of motors the overload relay moves toward closing its contacts every time the motor is started, and the timing preferably is adjusted originally by trial to be just great enough so that tripping is not completed. The timing thus determined experimentally, while the desired pickup value for the relay is not changed, then is applied to all other similar circuits.

The frequent movement of the moving element of the relay, during each starting period, is felt to be an advantage in preventing any tendency to stickiness and insuring that the relays remain in operating condition. After the original calibration and conditioning following installation, recalibration checks are required infrequently since operating conditions can be proved roughly by simply observing the amount of contact travel during the starting up of equipment. Actually, subsequent checks have never shown an appreciable divergence from the original values.

General Conclusions. The conclusions reached are not new and only agree with the general idea of providing instantaneous tripping devices to operate on fault currents plus low-pickup-value inverse-time relays for satisfactory clearing of prolonged overloads or low current internal winding faults.

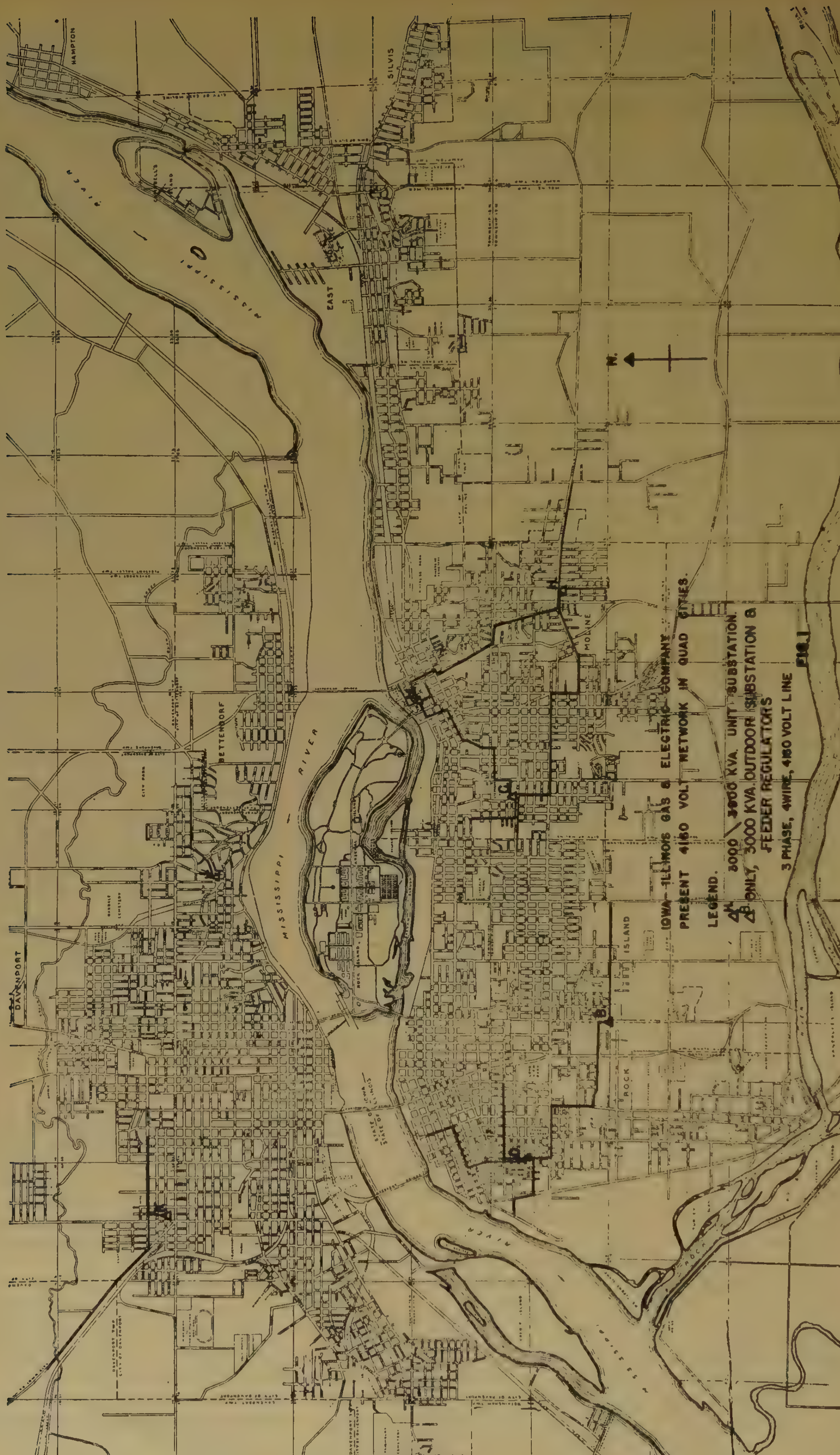

Figure 1. Present 4,160-volt 3-phase 4-wire network in quad cities

Δ^H—3,000/3,900-kva unit substation
Δ^B—only 3,000-kva outdoor substation and feeder regulators

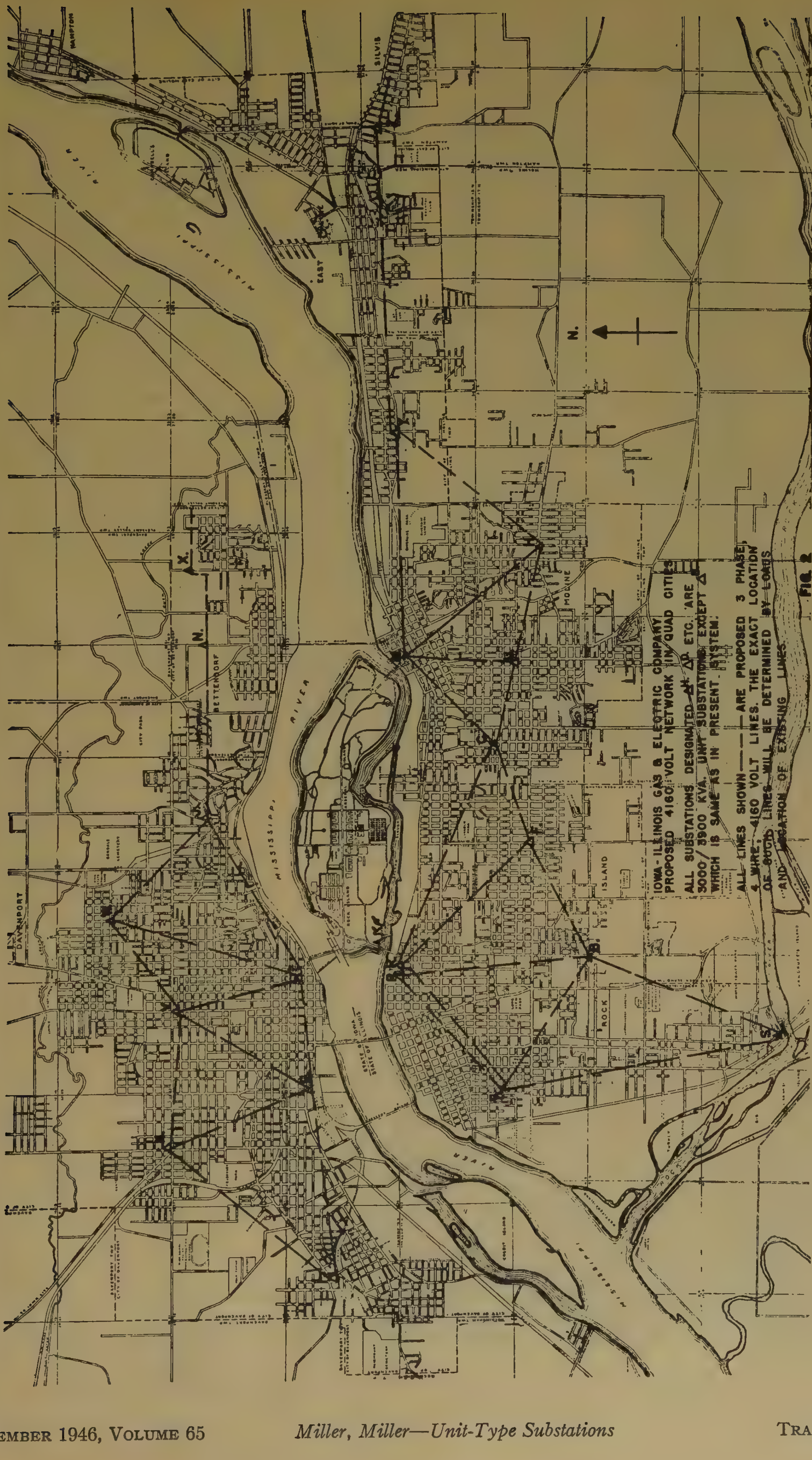

Figure 2. Proposed 4,160-volt 3-phase 4-wire network in quad cities

All substations designated Δ^H, Δ^D, and so forth, are 3,000/3,900-kva unit substations, except Δ^B which is the same as in the present system

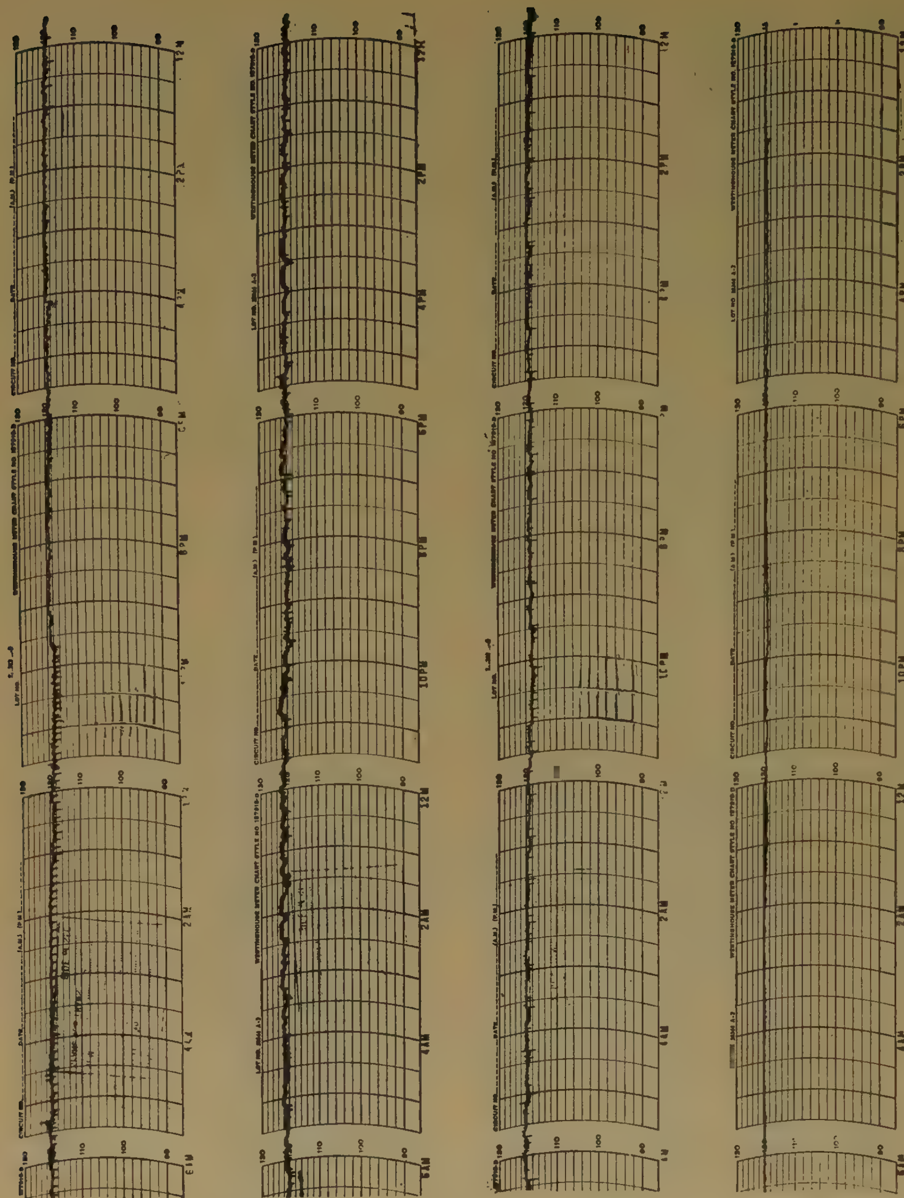

Figure 3. Typical voltage charts taken from secondary circuit approximately three spans from transformer

Transformers are located approximately one mile from the unit substation at center of 4-kv tie

Increased loads on existing substations required that additional larger capacity supply lines be constructed to those substations. The increased line capacity supplied through the larger number of step-down transformers raised the short-circuit duty on all existing station equipment. As a result much of the present station equipment became inadequate in interrupting capacity and thermal capacity. In addition, the fact that much of the existing station equipment was 2-phase also made it obsolete.

In 1939 it was decided to modernize the distribution system. After careful study of various types of distribution systems, it was decided that 4-kv primary networks would be instituted independently for the communities on each side of the Mississippi River. Because of the width of the river, it was deemed not advisable to join the 4-kv networks on opposite sides of the river. Involved in the study of the type of distribution system to be constructed was the matter of the type of substation to be utilized with the distribution system. After considerable study it was decided to use factory-built unit-type outdoor substations, these substations to be supplied from the 13-kv system and to be 3-phase 4-kv on the output side.

The majority of these substations will serve the residential areas, outlying commercial and small industrial areas. In general, no industrial customers will be served whose original demand exceeds 300 kva. In the areas where 13-kv circu are available as well as 4-kv circuits, indu trial customers having demands in exce of 100 kva will be supplied from the 13-l system.

The type of substation decided on h the following description: 3,000/3,900 kv equipped with transformer seconda breaker, four 3-phase 4-kv feeder compar ments, and necessary auxiliary compar ments for metering equipment, operatin transformers, and other essential equi ment. The individual feeders are equipp with overcurrent relays having instantan ous attachments. Three shot reclosers a used on each feeder, automatic reclosi being set for 15, 30, and 45 seconds wi lockout on the fourth outage. All su stations are equipped with air circu breakers, the original two substatio being equipped with stationary breake and group operated disconnecting switche The four later substations are equipp with drawout circuit breakers. The tran former secondary breaker, in addition being equipped with overcurrent relayi for bus fault protection and transform failure protection, is also equipped wi network and phasing relays which will co nect the 4-kv bus to the transformer a to the 13-kv circuit any time voltage proper magnitude and proper phase qua ity is applied to the 13-kv circuit. Indivi ual breakers are rated as 600 ampere 7.5 kv with an interrupting rating 150,000 kva.

Originally, two substations were pu chased in 1939, one of which had a cab entrance on the 13-kv side, the other ha ing an overhead entrance through cov bushings with lightning arresters mount directly ont he transformer cover. Exper ence with bushing flashover on this su station within the first month after it w placed in service caused a decision to reached that all future substations wou have 13-kv cable entrance. No switchi equipment is provided on the 13-kv si except a pole-top switch on one of the la poles before the 13-kv circuit goes unde ground.

The accompanying diagram, Figure shows substation locations and indicat the portions of the primary network whi so far have been tied together. The u substations are supplied either by a radi 13-kv feeder from one of the attended su stations or generating stations, or from tap onto an existing 13-kv tie line. T 13-kv tie lines make up a part of the ne work of that voltage embracing the gen ating stations and all attended substatio of the quad-city area. Only one unit su station is connected onto either a tie li or a radial feeder. The secondary side

each substation is tied into the 4-kv primary network. Each substation has at least two of its feeders, and ultimately will have all of its feeders, tied into other substations in the network. Radial load in the area through which each 4-kv tie line passes is balanced on the three phases of the tie line. This includes both the residential as well as the small commercial and small industrial load which already have been mentioned.

Actual construction of the primary network was delayed by the war. In addition to the substations now in service, one other substation has been on order since May 1945, and we contemplate the installation of six additional unit-type substations in the next 5-year period. The accompanying diagram, Figure 2, shows the ultimate network as planned with the future substations shown dotted. Substations are located in the small commercial areas adjoining residential districts. Where possible under zoning ordinances, substations will be located in the residential area. All 4-kv feeders are taken out underground, the actual underground distance being determined by the varying conditions at each substation site. The 4-kv tie circuits between unit substations are constructed as follows: outgoing riser cables with a 300,000-circular-mil single conductor and either paper-insulated lead-covered, or rubber-insulated with a Neoprene jacket. Where paper-insulated lead-covered cables are used and the neutral conductor is pulled in the same duct with the phase conductors, insulated wire is used. Generally, this has been a copper conductor of 4/0 size with a single layer of paper insulation and a lead sheath, although we have used some rubber-insulated braid-covered wire when wartime restrictions prohibited the use of lead on conductors insulated for less than 2,500 volts. The overhead conductors consist of three 4/0 phase conductors and a 1/0 neutral. The neutral wire is marked by means of white porcelain insulators and the phase conductors are always placed in sequence with the neutral as a marker. Construction, in general, is made to conform with the surrounding neighborhood. In some cases underground feeders have been run as far as two city blocks before rising to meet the overhead system.

Spacing of unit-type substations is determined by load in areas to be served, length of tie lines to other unit substations, planned future additions to the network, and by regulation based on the fact that bus regulation is used. Because the bus is regulated instead of the individual feeders, we have determined that no ties between stations will be over $2^1/_2$ miles long. In general, substations are spaced from 2 to $2^1/_4$ miles apart, and voltage regulation has been very satisfactory as shown by the accompanying charts, Figure 3.

Maintenance on the substations has been very slight. A regular weekly visit is made to each substation at which time all equipment is inspected, dust and dirt are removed by means of an electric vacuum cleaner and blower combination, records are made of kilowatt-hours and demand, and readings of the various operation counters are noted. A weekly inspection report is made out on each substation covering the various data taken, and space is provided for notes covering unusual maintenance, record of relay operations as indicated by targets and other matters of interest. A single spare circuit breaker is on hand which can be used to replace any of the existing breakers which require maintenance. For the circuits which are supplied from both ends, such as ties between substations, it is a simple matter to change circuit breakers without having interruption of service. For circuits which are strictly radial feeders, a time is chosen to interrupt service while the breaker is being changed which will cause the least inconvenience to the customer involved. In the six years of operation with these substations, only two minor cases of repairs in circuit breakers have been necessary.

We believe that the primary network

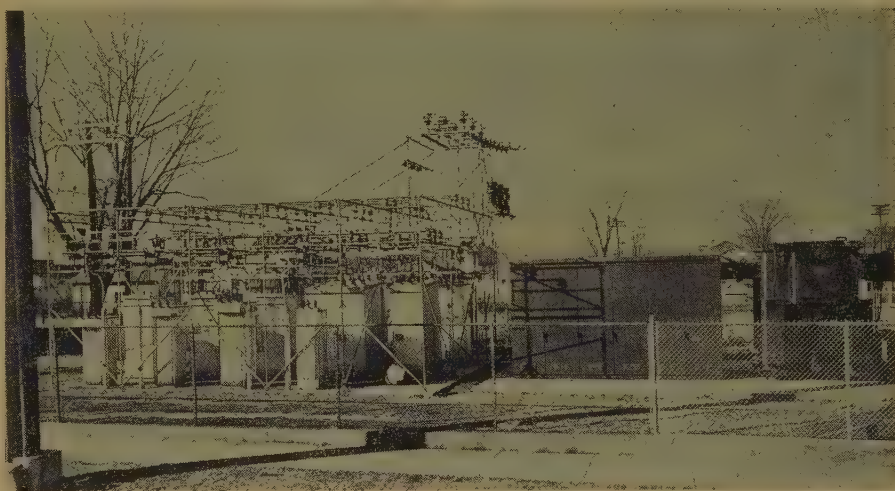

Figure 4. A 3,000/3,900-kva unit substation located in the same lot with a 1,500-kva substation of the open conventional design

The 1,500-kva open type substation was later removed and the material used at other locations

Figure 5. Close-up view of one of the standard 3,000/3,900-kva unit substations

and the use of unit substations have a number of advantages over the radial type of system. The location of the substation close to the load center keeps losses to a low value and provides for better regulation. The bulk of the energy is supplied at the subtransmission voltage which insures better system regulation and a more economical source of energy. The 3-phase low voltage lines provide better regulation with lower losses. The use of 3-phase tie circuits makes it easy to supply polyphase service to small commercial areas and outlying industrial locations. Spare transformer capacity is not required as in the conventional radial system. Voltage charts taken on the substation busses and on the customers' premises indicate that for the most part a smaller variation in voltage is maintained than where induction voltage regulators are used on radial-type feeder circuits. This of course accounts for better operation of appliances, particularly fluorescent lighting units, greater customer satisfaction, and increased revenue. Continuity of service through the use of primary network is excellent. The absence of induction feeder voltage regulators lowers maintenance costs as well as first costs, and operating and maintenance costs have been much lower than they are for the identical radial system.

We have found that a standard design of unit substation can be utilized to work out in the quad-city area as well as in the smaller outlying districts which incorporate a 3,000/3,900-kva transformer with one transformer secondary breaker and four feeder breakers. This size of transformer was chosen on the basis of cost per kilovolt-ampere and load density which we encounter in the area served, coupled with a reasonable distance between substations. Air blast equipment is supplied on the transformers when they are purchased and they provide an economical reserve capacity which is required occasionally in the event of other substations in the network being out of service at times of peak load.

Figure 6. View of substation with air circuit breaker in drawn out position

Figure 7. View of one of the units set in edge of residential area, showing landscaping used

In addition to the relays and other control equipment already mentioned as being standard with the substation used, we use a recording wattmeter, a recording varmeter, three recording ammeters, and a recording voltmeter of the strip chart variety to give us a continuous record of those quantities. These instruments are all connected into the leads of the transformer secondary so that the quantities measured pertain to the substation as a whole. A voltmeter switch is used to get a record of individual phase to ground voltages. A 1-week record per phase to ground quantity is normally obtained. No indicating or recording instruments are provided on the individual feeders. Test blocks are provided, however, on the control panel for each feeder to permit easy installation of portable recording instruments where checks need to be made for magnitude and division of load on each circuit.

Provisions for future expansion are excellent. It is possible to add additional unit substations at any time at the points of maximum load growth and it is also possible to move a substation in its entirety if load shifts dictate, thereby salvaging the bulk of the investment in a substation installation. System planning through the use of a-c network calculators has permitted the selection of the most economical size of unit-type substation, the most economical number of circuits, proper loading of the various substations, proper control of regulation for the complete distribution system through the use of bus regulated substations, and adequate provision for future growth.

Frequent use is made of the a-c network calculator to keep ahead of network growth, to anticipate the requirements for additional substations, and to check the operation of the existing network. In this way we are able to check substation locations with respect to load growth and to anticipate well in advance the locations where future substations must be installed, and to plan network changes and additions to meet system requirements in the best way.

Systems Development of Military Communications

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Synopsis: It is the purpose of this paper to set forth the broad phases of the problems involved in establishing a telephone and telegraph landline system in a military theater of operations, the manner in which these problems were met, and the results obtained in the North African Theater of Operations. The aspects reviewed are those involving the unification of the theater-wide wire facilities into a single system including maximum integration of civil plant and military installations, with centralized supervisory control of the available resources.

IN the North African Theater of Operations the landline communications system handled thousands of telephone calls and thousands of telegraph messages daily, over many hundreds of miles of telephone and telegraph channels involving distances, in some cases, greater than the distance between New York and Chicago. To provide an adequate system of communications for the Armed Forces in the field requires the fullest exploitation of the available communications resources, both personnel and equipment, if satisfactory service is to be furnished with maximum economy in the employment of these resources. The system communications problem involves a landline network and radio and submarine cable circuits operated as links in the landline network, both telephone and telegraph. To accomplish the mission of providing these communications, it is essential that an appreciation of the system philosophy be acquired by all concerned with the planning, engineering, installation, and operation of these facilities.

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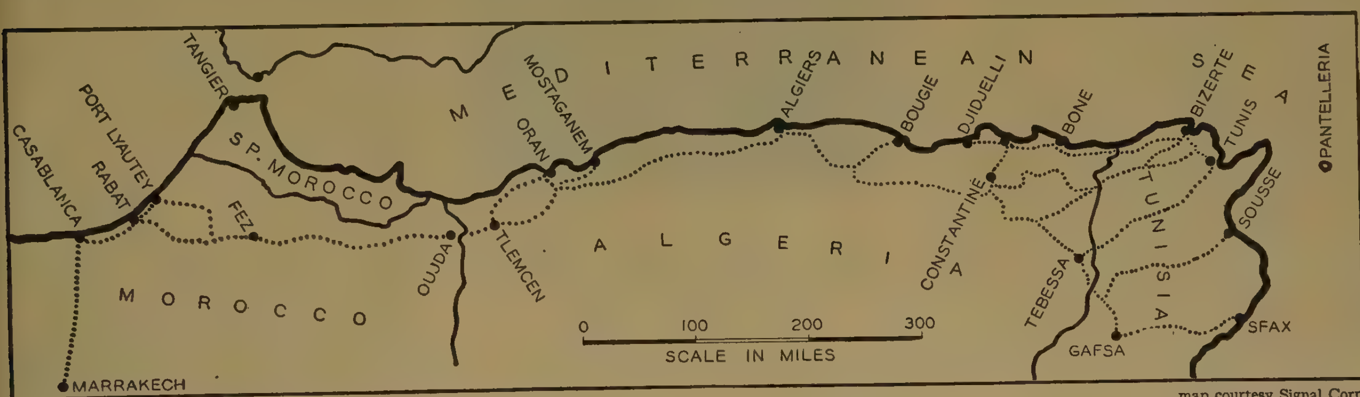
It might be worth while to consider for the moment the basic nature of the communication service required by the Armed Forces in areas in which extensive land and land-based air operations are carried out. In smaller units, perhaps up to and including division level in the ground forces, the primary purpose of communications is for fire control. That is to say, communications are required between commands posts, forward observers, observation posts, gun and infantry positions, and armored elements, to bring fire power to bear on the enemy when and where needed. Communications for this purpose have been largely on a point to point basis, especially in fast moving situations. However, as a given situation tends to become stabilized, a larger share of the traffic over the circuits is for administrative and supply service purposes, the number of users increases, the number of points required to be reached by telephone or telegraph increases, and switching facilities and system operation become a necessity.

Above division or perhaps corps level, the physical plant required begins to take on the appearance of a system network rather than a series of point to point routes. Many main routes are required, oftentimes passing through the areas of control of several commanders. For example, the requirements of the air forces for long haul circuits for intercept, tactical, and strategic air operations are governed by the deployment of the air forces. The location of the air installations depends not only upon the current tactical situation, but broad geographical and other considerations as well. Other governing factors in the location of main routes in the rear areas are the locations of ports of supply and other large administrative and supply centers. For these reasons the communications required to serve these areas must be considered as a system with many of the same problems we find in commercial operations.

In this connection, it seems desirable to look into the experience of commercial communications organizations and certain inherent features of their operations. The commercial operation of a landline communication service and its associated radio and submarine cable links has been described as a natural monopoly for the reason that the greater the number of facilities integrated into a single system, the greater is the value of the service to the users. It is on this basis that the need for a strong centralized control of the over-all operation is primarily recognized. This same necessity and reason exists in the development of a system of military communications. In addition, only by means of an integrated communication organization can the required flexibility in allocation of resources be attained to provide facilities when and where needed.

A major difference between the opera-



map courtesy Signal Corps

Figure 1. Map showing main landline routes in French North Africa

Approximate route distances, miles:

Algiers–Oran	340
Algiers–Casablanca	890
Algiers–Constantine	320
Algiers–Bizerte	580
Casablanca–Marrakech	150
Casablanca–Bizerte	1,470

tion of a military communication system in the field and commercial operations, however, is the method by which the requirements are limited with respect to the available supply of facilities. In general, the normal peacetime demands on a commercial service are kept in reasonable balance with the available supply of facilities by virtue of the cost of the service to the users, which in turn is related to the investment. This means of regulating the needs of military users, however, is not applicable in areas of military operations, which is another reason for centralizing appropriate technical control of communications at the highest level of military authority in the area served by the system.

Another major difference between military and commercial systems is the fact that the development of a commercial system is usually characterized by an orderly change of requirements, with time factors permitting the installation of an economical, well-designed plant. However, the pressure of limited time in the development of a military network in a theater of operations seldom permits the thorough co-ordination of all concerned to the extent desired.

System Philosophy

The need for centralized control of communications facilities becomes apparent when we realize how closely the activities of a communications installation at one location must be co-ordinated with those of the installations at other locations. Satisfactory electrical communication has been established between two points only when a conversation can be held between the two, or a message can be transmitted from one point to the other with such quality and dispatch as will satisfy the user. Delays due to unwarranted interruptions or lack of co-operation between those responsible for operations at various points in the system are evidence of unsatisfactory service.

It is not sufficient that all individuals and organizations place and splice fie wire or lead-covered cable; constru and maintain pole lines; place open wir install and line up carrier terminal ar repeater equipment; install, maintai and operate telephone and telegrap switchboards and radio installations i accordance with the same manuals of in structions. These activities must b carried out by organizations and person nel in several commands at the sam time. The only way this possibly can b done without delay and confusion is b adopting basic principles of military sig nal communication organization whic will permit the designation of a singl agency in advance to be responsible fo laying down the ground rules under whic this work is to be accomplished. Further more, these ground rules must be bindin on all commands whose communication are an integral part of the system as whole.

It may be accepted as axiomatic tha rarely will sufficient resources be availabl in a theater of war to satisfy the commu nications requests of all users. It there fore becomes incumbent upon those re sponsible for providing this service t exercise the utmost in judgment and skil to bring about maximum economy in th employment of these resources. In view of the technical aspects involved in th interconnection and interoperation of th various equipment located hundreds o miles apart in many cases, it is easy t see why the centralized control of tech nical administration by an experience and highly specialized staff assumes ar important role in systems development.

To consider that it is only necessary t provide communications within the chair of command, that is, from superior to sub ordinate commanders, is to fail to appre-

Signal Corps photo

Figure 2. Four-pin crossarm constructed by the Signal Corps on French civil open wire lead during combat in Tunisia

Figure 3. Signal Corps personnel operating a French civil switchboard with French interpreters at Oran

Signal Corps photo

Figure 4. Signal Corps personnel operatin Creed teleprinters (British) at Allied Force Headquarters in Algiers

Signal Corps phot

ciate the realities of modern operations. Were this the case, the provision of communications would be relatively simple. The problem is to provide a single network for use of all commanders and staffs of all elements of the armed forces occupying a given area, and it is under these conditions that complexity is added to the system. A further complication is added and in no small degree, in the provision of communications for joint operations involving the land, sea, and air forces of not only our own, but allied nations as well, utilizing the native resources of each, including civil systems.

In those overseas theaters in which organizational difficulties were encountered in systems development of communications during World War II, these difficulties stemmed from a basic deficiency in signal communications doctrine as promulgated by the War Department. Briefly stated, this doctrine sets forth the principle that every commander controls the system of communication required for his command. This principle of control by the several subordinate commanders must not be confused with the concept that the technical control of the single system of physical plant must be exercised by the common superior of these commanders. These two concepts of control are complementary and not inconsistent. For this reason they must be treated coequally in any pronouncement of signal communication doctrine for the Armed Forces.

Experience in the North African Theater of Operations—May to November 1943

Following the surrender of the Axis land forces in Tunisia in May 1943, the landline communication system of the Allies extended from Marrakech and Casablanca to Tunis and Bizerte in French North Africa, covering a maximum distance of about 1,600 miles. Communications operations expanded considerably following the cessation of ground combat operations for the mounting and support of two major amphibious operations, Sicily in July, and Italy (Salerno) in September 1943. To give some idea of the magnitude of the facilities required, following is a tabulation of the long haul channel mileages operated during the fall of 1943:

1943	Channel Miles: Telephone	Channel Miles: Telegraph
September	26,500	29,300
November	38,300	34,400

The outside plant making up this network consisted of existing French civil (Services des Postes des Telegraphes et des Telephones) open wire leads and buried cable, American 10-pin and 8-pin open wire leads, American spiral four cable, and buried cable placed by both the American and British forces. Many of the French civil open wire leads in Tunisia had to be rehabilitated following combat operations in that area.

The equipment used to derive the long haul channels included French civil systems and varieties of American and British equipment available at the time. The French installations consisted chiefly of 1- and 3-channel telephone carrier systems operated over the open wire leads between Rabat and Oran. American telephone carrier equipment consisted of types *H* and *C* Western Electric systems, providing one and three carrier frequency channels, respectively; and Signal Corps type CF-1 which provides one voice frequency telephone channel and three carrier frequency telephone channels. British equipment consisted of the so-called "one plus one" (one voice frequency telephone channel plus one carrier-frequency telephone channel) and "one plus four" (one voice frequency telephone channel plus four carrier frequency telephone channels) tactical systems, and British SOS and SOT (three carrier frequency telephone channels) commercial systems.

Both American and British voice frequency telegraph systems were employed for long haul telegraph (teletypewriter and teleprinter) use. American types consisted of type 40-C-1 Western Electric (12 channels) equipment and the Signal Corps CF-2 (4 channels) tactical system. British voice frequency telegraph systems used, were the "speech plus simplex," "speech plus duplex," and "6 channel duplex" tactical systems. The "speech plus duplex" (S plus Dx) and "speech plus simplex" (S plus Sx) systems are designed to derive carrier telegraph from a portion of the frequency band used by a telephone channel while retaining the use of the channel for speech transmission. In the S plus Dx system the band from about 1,500 to 2,000 cycles is eliminated from the speech transmission circuit and used for telegraph purposes. The carrier midband frequencies employed are 1,680 cycles for one direction of transmission and 1,860 cycles for the other direction. Carrier is transmitted for marking and interrupted for spacing. The S plus Sx system provides service in both directions on a circuit but in only one direction at a time (American half

Signal Corps photo

Figure 6. Main distributing frame of the Freedom Exchange installed by the Signal Corps at Algiers

Signal Corps photo

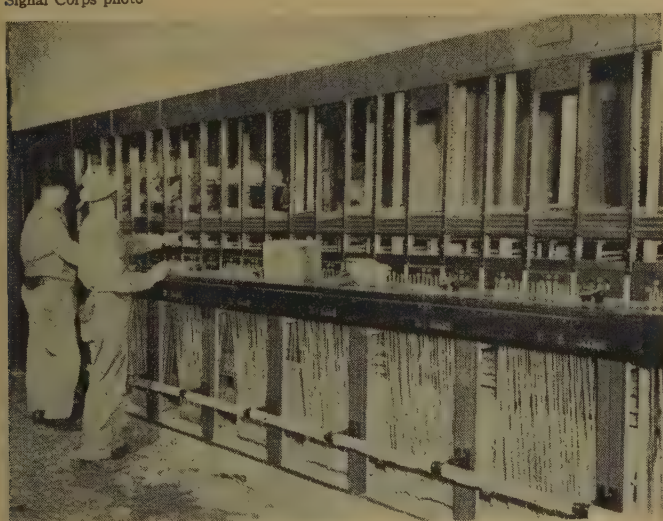


Figure 5. Switchboard installers of the Signal Corps at work on an 8-position Western Electric 11 type switchboard at Algiers

Duplex). A choice of carrier frequencies is available in this latter system depending on which of the three types of the equipment is used. Numbers 1 and 2 are arranged to use frequencies of 300, 900 or 2,300 cycles per second. The number 3 equipment uses frequencies of 300, 1,740, or 2,300 cycles per second.

Multichannel very high frequency radio equipment as developed and later used in other overseas military areas as links in landline systems was not available in the North African Theater; and a discussion of the point to point radio operations is outside the scope of this paper. In the fall of 1943 a limited amount of broadcasting was done over transmitters in the vicinity of Tunis from studios in Algiers, using landline program supply channels. Although these channels were not high quality, they did serve the purpose. Musical programs of the Special Services and foreign language news were broadcast for short periods at night. Also, on several occasions news broadcasts picked up at points in Tunisia were carried over landlines to Algiers where they were transmitted by radio to this country for broadcast.

The Mediterranean area has a rather well-developed civil network of submarine cables which proved to be extremely useful to the Allied Forces. A considerable volume of intratheater telegraph traffic was carried over these circuits particularly after the islands in the Mediterranean between French North Africa and Italy were occupied. In respect to security considerations, submarine cable circuits are unique in that they are acceptable for transmission of military telegraph traffic in clear text, including messages classified as secret, providing the water areas through which the cables pass are in complete control of friendly forces. In view of the shortage of automatic cryptographic equipment, the advantage of this was that arrangements could be made if the condition just mentioned were fulfilled, for the clearing of large volumes of telegraph traffic without the time-consuming burdens of manual cryptography.

To develop a theater-wide system of communications involving complex operating features as just described, it soon was found that staff responsibility without technical administrative authority over the operating personnel of the various subordinate commands was entirely unsatisfactory. The solution to this problem in French North Africa was to activate liaison offices called long lines district offices, in Oran, Algiers, Constantine, and Bizerte to be under the direct control of the wire division of the signal section, Allied Force Headquarters. This action was taken in September 1943. Signal Corps officers and enlisted men were assigned to these offices who were entirely familiar with the "big picture" of the system and who were technically competent to analyze maintenance and operations results.

Figure 7. Signal Corps personnel operating an 8-position Stromberg-Carlson 18 type switchboard at Constantine

Signal Corps photo

Telephone Operations

In September 1943 it was found necessary to restrict the number of telephones having access to the long distance service. Long distance calls were permitted from about 50 per cent of all the telephones connected to military switchboards. The effect of this policy was an immediate improvement in the service, and by October between 85 and 90 per cent of all long distance calls which were placed were completed on the first attempt on a non-hang-up basis.

The principal headquarters telephone switchboards completed an average of approximately 69,000 calls per day in the month of November 1943. Conservatively estimating the number of calls handled in addition on smaller switchboards for which records were not available, the total number of telephone connections was probably in excess of 150,000 per day.

Table I

Location and Switchboard Designation	November 1943		
	Type of Board*	Number of Positions	Number of Calls Per Day
Algiers			
Algiers long distance	WE 11	6	5,570
Freedom (local)	WE 11	10	20,457
Bizerte			
Bizerte (combination local and long distance)	TC-1	5	5,557
Eagle (local Eastern Base Section)	TC-1	3	5,003
Casablanca			
Beacon (combination Atlantic Base Section)	WE 550	3	6,589
Constantine			
Constantine (combination)	SC 18	8	5,829
Oran			
Merit long distance	TC-1	6	3,675
Merit local (Mediterranean Base Section)	WE 11	8	16,086
		Total	68,766

* WE—Western Electric; TC-1—Signal Corps nomenclature; SC—Stromberg-Carlson.

Table II. Landline Telegraph (Teletype) Traffic Data* for Certain of the Major Signal Centers in French North Africa—June and November 1943

Month	Allied Force Headquarters	Atlantic Base Section	Mediterranean Base Section	Eastern Base Section
June	95,638	27,379	42 938	18,650
November	114,057	9,886	52,187	27,926

* Daily average—incoming and outgoing words (excluding press).

The record of calls passed on the principal boards as recorded from actual peg counts is shown in Table I.

Telegraph Operations

The first voice frequency telegraph system in French North Africa using American equipment was installed in March 1943. Signal Corps CF-2 four channel equipment was installed in Algiers and Oran and operated over a French civil underground cable circuit. Prior to that time all voice frequency telegraph channels were provided by means of British equipment of the type mentioned earlier. Also, in the same month of March, sufficient teletype switchboards (Signal Corps BD-100) had become available to establish a common user telegraph printer network. Switchboards of this network were installed at Oran (Mediterranean base section), Algiers (Allied Force Headquarters, Air Force, port, and other service organizations), and at Constantine (Eastern Base Section); and at later dates the system was extended to include Casablanca, Tunis, and Bizerte.

The telegraph traffic load of several of the larger headquarters in French North Africa for June and November 1943 is shown in Table II.

Technical Administration

The development of a network of communications into an efficient system requires a scheme of technical administration which will enable each member of the communication team to know not only what his own duties and responsibilities are to carry out a given function, but also what other members of the team at distant points should be doing. For routine matters this is best accomplished through the medium of a set of uniform practices or instructions issued under a competent order to all concerned. A few of the activities which must be covered in this way are service order and trouble reporting routines, and long distance circuit order and facility record practices. These instructions must take into account all the conditions affecting the operation of the communication system, such as the communication procedures of all the military and civil agencies and commands involved in the operation, and the numerous types of equipment and organizations available to the military forces and civil communication administration. Instructions of this kind are applicable only to the communication system of a particular area of operations, and for this reason they must be prepared in the field by the agency responsible for the operation of the system. In the North African Theater of Operations these instructions were issued as "Operating and Maintenance Instructions," by the wire division of the signal section, Allied Force Headquarters.

By the fall of 1943, plant operations showed considerable improvement over earlier results. Contributing largely to this improvement were the extensive inside and outside plant preventive maintenance programs which were put into effect during the summer, the introduction of universal patching facilities, and the substantial on-the-job training acquired by the maintenance personnel. In addition, of major importance was the improved operation of the plant on a system basis following the activation of the afore-mentioned long lines district offices. For example, during the two months immediately following the establishment of these offices, telephone carrier troubles of one hour or more duration decreased 40 per cent, and similar telegraph carrier troubles decreased 50 per cent during the same period. These improvements in carrier maintenance operations provide strong evidence as to the soundness of the system doctrine applied to military communications operation. The over-all experience in the operation of the landline system in French North Africa leads inevitably to the conclusion that a single chain of command must exist from the very top down through all signal communication organizations involved in providing a communications system for military operations in a theater of war.

Frequency Compensated A-C Ammeters and Voltmeters

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Synopsis: The increased use of a-c instruments over a frequency range of 25 to 3,000 cycles for aircraft and industrial applications has imposed new requirements on instrument design. It is recognized that the standard type of a-c ammeters and voltmeters normally calibrated for use at a given frequency are subject to errors in indication when used at other frequencies. This paper describes the manner in which one line of a-c ammeters and voltmeters can be compensated to have minimum errors resulting from changes in frequency over a range of 25 to 3,000 cycles. Errors found in the uncompensated and compensated types of the instruments with changes in frequency are shown, and a practical means of determining the amount of compensation and its application to ammeters and voltmeters is presented.

FOR many years the conventional types of a-c ammeters and voltmeters have been designed for operation over a small range of power frequencies from 25 to 125 cycles per second and have been used primarily on 25-, 50-, and 60-cycle circuits. In recent years, however, the operating frequency range has been extended[1] and instruments have been designed with improved magnetic materials to operate accurately over a range of 15 to 12,000 cycles. Some of the more common frequencies encountered in the extended range have been 180 cycles for portable machine tools and high speed motors, 400 and 800 cycles for aircraft, 960 to 3,000 cycles for electric furnace applications, and from 8,000 to 12,000 cycles for industrial heat treatment use.

The more general method of instrument adjustment has been to calibrate at 60 cycles for use at 25 to 125 cycles and at the specified operating frequency for higher ranges. This procedure in the past has proved satisfactory and, generally speaking, variations up to 20 per cent of the calibrated frequencies have not produced objectionable errors. A rapidly increasing demand, however, has developed recently for instruments to be used over much wider frequency ranges. For example, in aircraft there have been applications for instruments for use on not only 400 but 800, 1,200, 1,600, and 2,100 cycles. The problem of equipping these airplanes and maintaining replacement supplies would be simplified greatly if an instrument calibrated at one frequency

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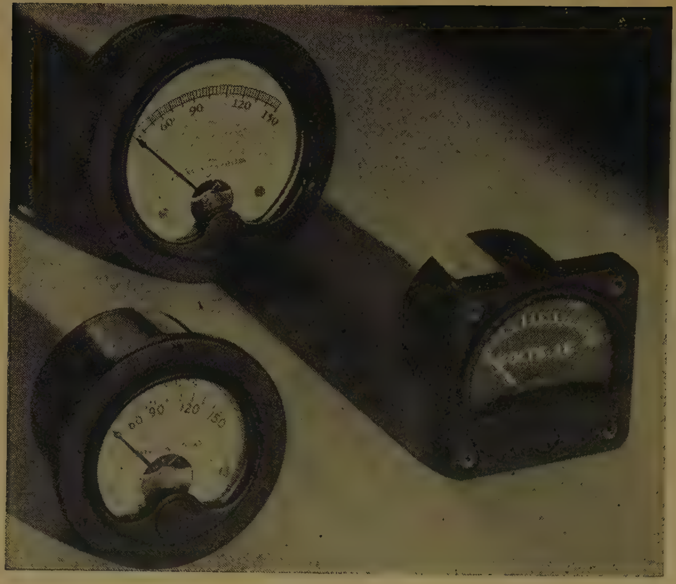

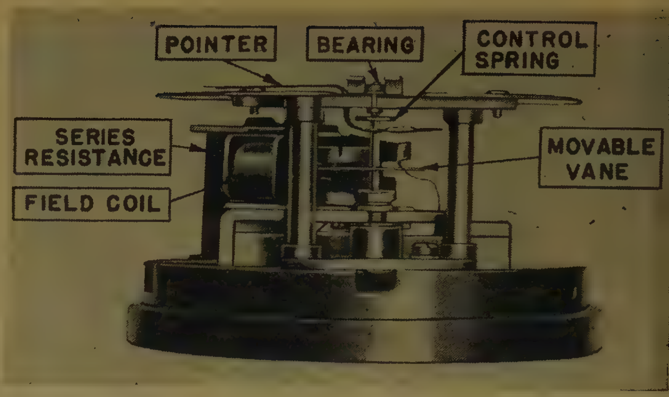

Figure 1 (left). Representative panel type a-c instruments for industrial and aircraft applications

Figure 2 (above). Instrument assembly showing mechanism and internal construction

also would perform accurately over the complete range. Similarly, in many modern industrial and electronic applications the operating frequency may vary over ranges much in excess of 20 per cent of the calibrating frequency. These conditions have created a need for instruments which are compensated over a wide frequency range. A method of compensation and the results obtained are described in this paper.

Errors in Uncompensated Voltmeters and Ammeters

The moving iron instrument of the attraction type used for many a-c ammeters and voltmeters is shown in Figure 1 in typical sizes. Briefly, the mechanism of this instrument consists of a vane of high permeability material moving within a fixed coil and mounted on a nonmagnetic shaft which carries the indicating pointer and is supported in suitable bearings. When current flows in the fixed coil, a magnetic field is set up which tends to pull the magnetic vane into the coil. The rate with which the vane is pulled into the coil is controlled by a flux distributor, and the movement is opposed by the torque of a control spring attached to the shaft. The internal construction of this type of instrument is shown in Figure 2.

This mechanism is used in ammeters by allowing the current to be measured to flow through a field coil. For voltage measurements, the mechanism is connected across the line and in series with a fixed resistance, to permit a current proportional to voltage to flow through the field coil. The electrical circuit of a standard uncompensated voltmeter is shown in Figure 3.

The voltmeter is subject to frequency errors which are caused by changes in the inductive reactance of the field coil and of the series resistor, and in the effective a-c resistance of the coil. The impedance of this typical design at 60 and at 3,000 cycles is shown vectorially in the two diagrams in Figure 4. While the change in inductive reactance is directly proportional to the change in frequency, the relation of the change in effective resistance with frequency is more complex. It has been shown previously[2] that the eddy current losses in magnetic circuits vary as a nonlinear function of the frequency, and have the same effect as a resistance which shunts the field coil and decreases the current through it. The combined effect of these two factors is to increase the circuit impedance and decrease the indication for a given voltage as the frequency increases.

An example of errors in an uncompensated voltmeter of this type is shown by the calibration check of a typical 150-volt voltmeter in Table I. The instrument was calibrated at 60 cycles and checked over a frequency range of 60 to 3,000 cycles. The differences in readings at 60 cycles and other designated frequencies are expressed in per cent of full scale voltage. This rating of voltmeter has a series resistance of 14,000 ohms which limits the current through the field coil to 10 milliamperes.

The frequency error in ammeters of the moving iron type is caused only by the change in eddy current losses and is not affected by inductive reactance changes which merely cause a change in the voltage drop across the coil, but not in the current through it. As a result, the overall error is considerably less than that found in voltmeters. For example, the error of a 5-ampere ammeter, similar in general design to the 150-volt voltmeter described and adjusted on 60 cycles, does not exceed minus 3 per cent at full scale indication on 3,000 cycles.

Method of Frequency Compensation for Voltmeters

A study of frequency errors shown in Table I, indicates that voltmeters should be compensated for frequency changes when calibrated at one frequency and used over a wide range of different frequencies. The ideal, from the users standpoint, would provide compensation to hold the magnitude of the total effective impedance constant even though the values of the resistive and reactive components change as the frequency varies. A study of prior methods[3,4] has shown that compensation can be obtained by shunting all of the series resistance with a capacitor of required value if the resistance of the series resistor is relatively large as compared with that of the field coil.

This method is very undesirable for use in quantity production, as the value of capacitance must be held within much closer limits than the standard plus or minus 20 per cent tolerance in the micro-

Table I. Uncompensated Instrument Errors Over Frequency Range

Expressed in Per Cent of Full Scale Value

Standard Voltage	Cycles 60	400	1,200	2,000	3,000
60	0	−0.8	−2.3	− 4.9	− 8.9
90	0	−1.1	−3.6	− 7.6	−11.2
120	0	−1.1	−4.7	− 9.3	−14.8
150	0	−1.5	−5.1	−12.0	−22.0

farad values of capacitors which are available in quantity, and in the dimensions required. Furthermore, such capacitors will be subject to peak voltages which are more than can safely be impressed on them. For example, a capacitor which shunts 14,000 ohms of a 15,000-ohm instrument circuit would be subjected to a peak voltage of 198 volts at full scale indication, and to even much higher values for intermittent overloads to which instruments may be subjected.

These difficulties, however, can be overcome largely by shunting only a portion of the series resistor and varying the resistance value of this portion to match the actual capacitance of a given capacitor. Since the value of this resistance is determined during the test procedure, and the resistor spool then is wound to the required resistance, it is not necessary to maintain a large inventory of these resistor spools, and commercial tolerance capacitors can be used without difficulty in quantity production. In addition, the voltages likely to occur across them are reduced to safe values.

With this method, the amount of compensation and, therefore, the magnitude of the remaining error in a given voltmeter, depends upon the per cent of the series resistor which is shunted and the value of capacitance used. The effect of different per cents of series resistor shunted, and the magnitude of errors over a 60- to 2,500-cycle frequency range, is indicated in Figure 5. As shown, shunting 58 per cent of the resistor with a 0.004-microfarad capacitor reduces the errors to a maximum of 1/4 per cent over the frequency range. If, however, this compensation is used on 3,000 cycles, a 1 per cent

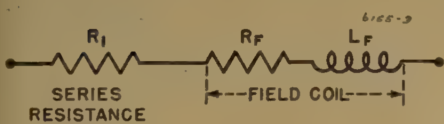

Figure 3. Schematic diagram for uncompensated moving iron type voltmeter

R_F—Field coil resistance
L_F—Field coil inductance
R_1—Series resistance

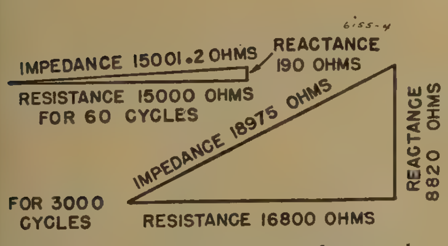

Figure 4. Vector diagrams showing how internal impedance varies with frequency

error will occur. On the other hand, errors over the range extended to 3,000 cycles can be made negligible by shunting 75 per cent of the resistor with 0.003 microfarad.

Application of Method to Actual Voltmeters

It is important to note that the preceding values are for an individual voltmeter and will differ slightly with each instrument as a result of the tolerances which must be allowed in the components. The different values of series resistance which were shunted by a fixed value of capacitance in order to fully compensate a group of six instruments are shown in Table II.

Table II. Variations in Resistance Shunted by Fixed Capacitance

Capacitor Rating, μf	Resistance Shunted, Ohms	Resistance Unshunted, Ohms	Per Cent Shunted
0.003	10,300	3,700	73.6
0.003	10,385	3,615	74.1
0.003	10,390	3,610	74.1
0.003	10,520	3,480	75.3
0.003	10,715	3,285	76.6
0.003	10,880	3,120	77.7

This method of shunting a portion of the series resistance approaches closely the ideal compensation by maintaining nearly constant the magnitude of the total effective impedance over wide frequency ranges. This is shown in Table III for which the equivalent resistance R_e and the equivalent reactance X_e have been computed from measured values for each component by using the following equations:

$$R_E = R_F + R_1 + \frac{R_2}{R_2^2\omega^2C^2+1} \text{ ohms}$$

$$X_E = j\omega\left(L_F - \frac{R_2^2C}{R_2^2\omega^2C^2+1}\right) \text{ ohms}$$

where R_F, L_F, R_1, R_2, and C refer to the components shown in Figure 6. The vector relationship of these values at 3,000 cycles is shown in Figure 7.

The data in Table III were taken for the full scale point only, and other points on the scale will vary slightly as a result of changes in the demagnetizing effects in the magnetic circuit resulting from the movement of the vane in and out of the field coil. The magnitude of such changes is shown in Table IV and is considered satisfactory for the applications previously mentioned. It is significant to note that the scale distribution changes with frequency and that such changes are more pronounced at the higher frequencies.

The application of this method of frequency compensation to large quantity production of voltmeters is entirely practical. The voltmeter first is adjusted at 60 cycles with the fixed coil in series with the resistance required for full scale. The frequency then is changed to 3,000 cycles, which increases the instrument impedance and decreases the pointer indication. A commercially available capacitor of approximately the right value then is connected across a sufficient amount of series resistance to bring the pointer again to the full scale point. This adjusting technique can be adapted easily to simplified test procedure by using standard resistors of different values which can be quickly connected to the instrument circuit by a selector switch.

Compensation of Lower Range Voltmeters

The discussion thus far has been restricted to voltmeters rated 0–150 volts. Lower range voltmeters can be compensated in the same manner if certain new problems are solved. In order to assure adequate factors of merit and temperature compensation, it is necessary to maintain a given number of ampere-turns for the field coil, and a definite ratio of the resistance of the copper in the

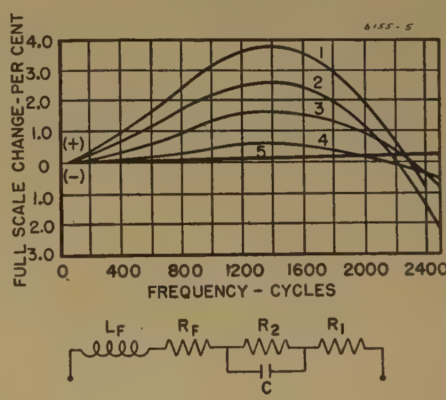

Figure 5. Degree of compensation for different percentages of series resistor shunted by specified value of capacitance

Decade Box For Resistors R_2 and R_1

Curve	Ohms R_F	R_2	R_1	μf C
1	1,000	3,645	10,650	0.025
2	1,000	3,300	10,700	0.025
3	1,000	4,200	9,800	0.015
4	1,000	5,750	8,250	0.008
5	1,000	8,100	5,900	0.004

field coil to that of the series resistor which is wound with negligible temperature coefficient material. These and other factors result in the series resistance decreasing more rapidly than the field coil impedance, as the full scale voltage rating is reduced. As a result, voltmeters in ranges below 75 volts do not have sufficient series resistance to be shunted for correct frequency compensation. It is necessary, therefore, either to make some sacrifice in performance or to design special high frequency instruments for such ratings to obtain effective compensation.

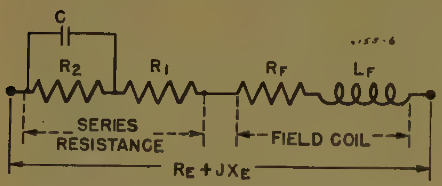

Figure 6. Schematic diagram for compensated moving iron type voltmeter

R_F—Field coil resistance
L_F—Field coil inductance
R_1—Unshunted series resistance
R_2—Shunted series resistance
C—Capacitor

Compensated Ammeters

It has been pointed out previously that the frequency errors in the ammeters are relatively small as compared with those found in voltmeters and principally are a result of eddy current losses. These losses can be reduced to a minimum by the use of high resistivity material such as phosphor-bronze for the field coil form and high permeability material such as permalloy, mumetal, and nicaloi in proper combination in the magnetic circuit. As a result, errors in ammeters are reduced to negligible values over the frequency range of 60 to 3,000 cycles.

Table III. Variation of Total Effective Impedance With Frequency

Frequency, Cycles Per Sec	R_E, Ohms	X_E, Ohms	Z_E, Ohms	Per Cent Variation of Z_E From Value at 60 Cycles*
60	15,030	70	15,030	0
400	15,090	493	15,090	+0.40
1,200	15,030	1,280	15,080	+0.33
2,000	14,820	2,130	14,970	−0.40
3,000	14,300	4,290	14,960	−0.50

* With the sign reversed these values are equal to the per cent error at full scale caused by frequency variation.

Table IV. Typical Scale Errors Over Frequency Range in Per Cent of Full Scale

Standard Voltage	Cycles 60	400	1,200	2,000	3,000
150	0	0	0	0	0
120	0	+0.1	+0.2	+0.6	+1.4
90	0	−0.3	−0.8	−0.1	+1.1
60	0	+0.4	−1.1	−1.1	−0.7

Conclusions

A practical technique of frequency compensation has been developed which can be applied easily to standard a-c instruments. The compensation has been accomplished first, by the use of improved magnetic materials in the ammeter and voltmeter, and secondly, by the addition of a commercially standard capacitor to the voltmeter.

By the application of these simple improvements in existing designs it is possible to use instruments normally adjusted for accuracy to within plus or minus two per cent of full scale value for one frequency, over a wide range of 25 to 3,000 cycles without incurring additional errors in excess of two per cent.

It is hoped that this work will encourage the use of frequency compensated a-c instruments to replace those now calibrated for a standardized frequency of either 60, 400, or 800 cycles as well as those calibrated for other specific frequencies. This will insure more accurate measurement of current and voltage over a wide range of frequencies.

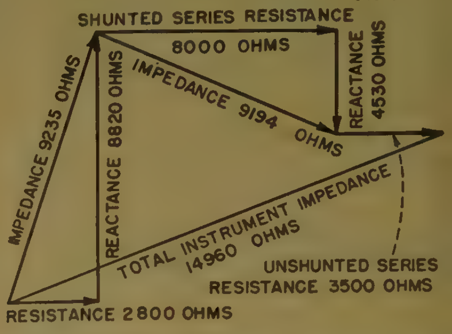

Figure 7. Vector diagram showing total instrument effective impedance at 3,000 cycles

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Sensitive Ground Relay Protection for Complex Distribution Circuits

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FELLOW AIEE

Synopsis: Protection engineers always have desired to obtain a system of ground fault protection that will discriminate and actually trip out only the ground faulted circuit of a complex distribution network. This has been a difficult chore because the magnitude of the ground current often varies so greatly as to be out of the range of the normal protective relay settings. A system is proposed here that not only has proper discrimination, but also operates on very low amounts of ground current with the maximum ground current held to such a low value that practically no harm or damage is done by the ground fault. This development was encouraged by the necessity of joint use of poles by lines of communication circuits and distribution circuits. This method comprises the installation of a small bank of grounding transformers at each of the stations where the lines originate or terminate. With these small ground banks the maximum ground fault current is very small. Adequate relays and systems are used to isolate only the faulted sections of line. The installation of one 11-kv and one 16.5-kv loop will be described. These two systems have had several years of successful operation.

Conclusions

THIS type of sensitive ground relay protection will discriminate on ground currents in the order of one per cent of the current transformer ratings, that is, with a 200–5 current transformer on the circuit, a fault of two amperes ground current will cause proper tripping.

The maximum ground current can be arranged to be from 40 to 60 amperes. This low current can do little harm as compared to the damage caused by a current of several thousand amperes that may occur in the solid grounded neutral systems. It has been found by years of operating experience that the damage done on ground faults with this type of protection is negligible.

Tests have been made with tree grounds and it has been shown that conductor contacts on green trees do not pass sufficient current to trip the lines out from this cause. Years of operating experience have confirmed this fact.

The operation of this relay scheme can be made very fast, thus eliminating ground faults rapidly. Since the control of the ground relays can be used to control the phase relays, phase-to-phase short circuits also can be cleared rapidly.

Introduction

Continuity of service to the customers of electric power utilities always has been one of the main problems of the protection engineer. Complex and interconnected high voltage circuits have been common for a long time. The protection of these circuits has been satisfactory and is improving continually. The lower voltage circuits have the same advantage of giving better continued service when interconnected. This has been done for some time but there has been no reliable way of isolating ground-faulted feeders correctly when the range of ground current has been so great. Therefore development of such a system was begun. Various standard types of relays were applied to many systems and no reliable discrimination could be obtained.

Then two further requirements became apparent:

1. If successful operation is to be obtained in joint pole use, the distribution line upon contacting the telephone circuit should be isolated rapidly and with as small amount of current as necessary to properly clear the distribution circuit. It is important to both the telephone utility and the power utility to make these fault damages a minimum as they both serve the same customers and continuity of both services is important.

2. The damage to property and the hazard caused by fallen live wires makes it necessary to isolate and de-energize them as rapidly as possible and with as small amount of ground current as necessary.

It can be seen that the same relay system can accomplish each type of ground fault isolation. During many years, ground tests were conducted with many types of relay systems and many improved ones were developed, but most of them were not adequate to service a complex distribution system. The system described here was finally suggested and tried. Although rather complicated for distribution circuits, it has proved successful with years of experience. Several improvements have been made to facilitate operation. The system herein described is now in service on some of the

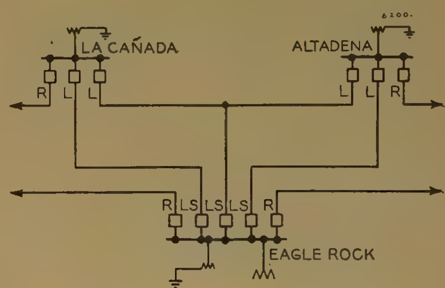
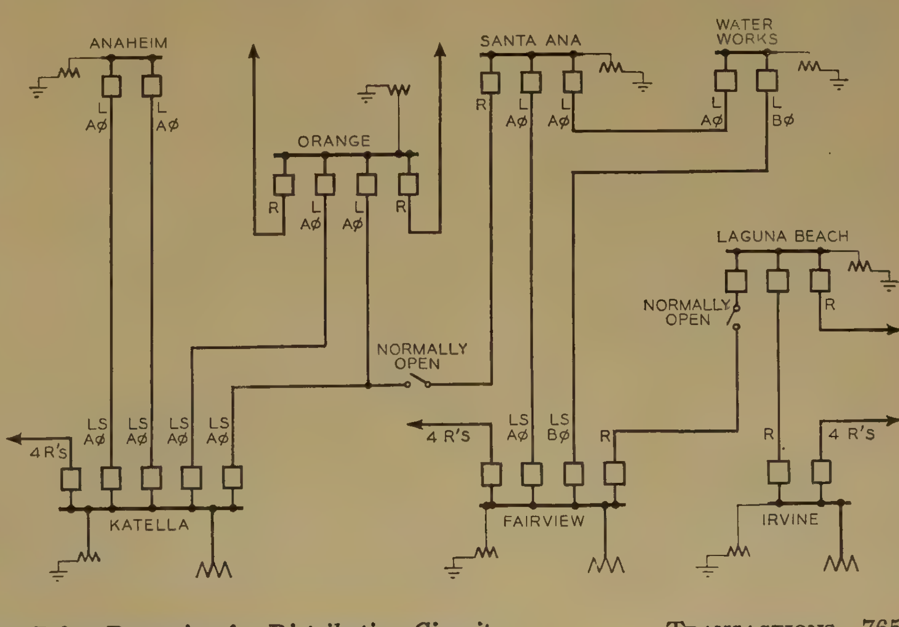


Figure 1. Eagle Rock substation 16-kv loop system

Figure 2 (below). Katella and Fairview substations 11-kv loop system



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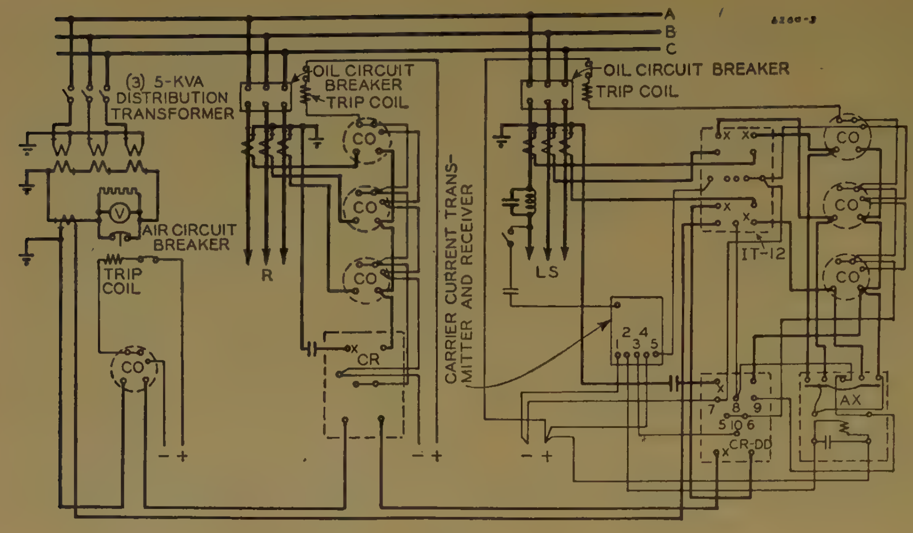

Figure 3. Source station relay system

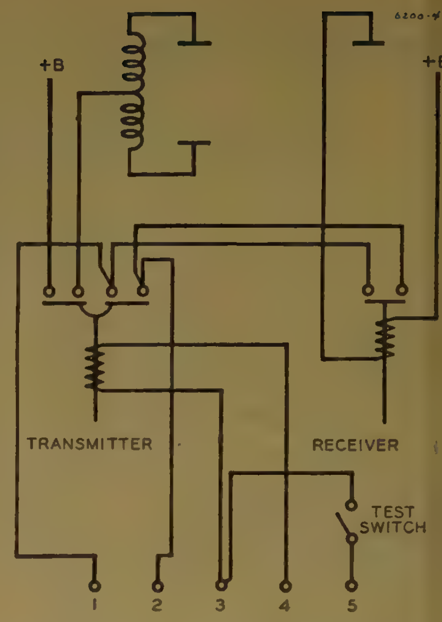

Figure 4. Carrier set relay system

distribution circuits of the Southern California Edison Company Ltd.

Complex Distribution Systems

The sensitive relay systems were applied to one 16-kv network as shown in Figure 1 and a more complicated 11-kv network as shown in Figure 2.

At each station a small grounding bank of three 5-kva distribution transformers is installed.

It can be seen that a separate type of protection is required for:

1. Radial feeders, designated by *R*.

2. Loop feeders at the source station, designated by *LS*.

3. Loop feeders at the receiving stations, designated by *L*.

Relay Systems

The diagram in Figure 3 shows the relay protection, type *R* and *LS* required for source stations such as Eagle Rock substation shown in Figure 1, and Katella and Fairview substations shown in Figure 2. This diagram shows only one radial feeder and one loop feeder. Additional feeders are exact duplicates of one of these types. Figure 4 is a schematic wiring diagram of the relays used in conjunction with the transmitting and receiving sets. Figure 5 is an internal diagram of the special fault detecting relay *IT-12*. The radial feeder relay system is the same as that described in a previous paper.

It can be seen (Figure 3) that the grounding bank consists of three 5-kva distribution transformers. The primary windings of these transformers used in Figure 1 are 16.5 kv, the secondary windings, 115 volts; and the ones used in Figure 2 are 12 kv to 120 volts.

The grounding bank secondary delta is closed by an air circuit breaker which is controlled by an over-current relay. In the event the relay protection takes longer than a predetermined time to clear a ground fault, the over-current relay will operate, thus opening this delta circuit breaker. When the circuit breaker is open a voltmeter across the delta indicates percentage ground faults. Only at the source station is there a resistor of such conductivity as to prevent resonance, thus preventing high voltages when the system becomes ungrounded.

Because the grounding banks at the various stations are small, their impedance is high in relation to the line impedance. Therefore a ground fault at any location will cause all grounding banks to produce almost the same amount of ground current. The maximum ground current depends upon the number of ground banks used in a given system when a solid ground fault occurs. The magnitude of ground fault current then will range from a minimum to this maximum, inversely as the resistance in the ground fault.

Since the line impedance is so low, the problem of discriminating by time is impossible. It therefore is necessary to use a blocking method to prevent improper ground relay action. In the two locations shown in Figures 1 and 2, carrier current sets[2] are used to send the necessary lockout signals. Pilot wires could be used for the signal channel but at the above locations the carrier sets proved more economical.

The directional ground relay type *CR-DD* is made as shown by the internal connection diagram, Figure 6. This relay is similar to a directional-controlled single-phase power directional relay except that the directional contacts have duodirectional function. When contact 8 closes to contact 10, it activates a transmitting relay sending a signal to the remote end of that circuit to lock out the tripping relays, thus preventing them from operating. When contact 8 is closed in the opposite direction, the relay is in tripping condition.

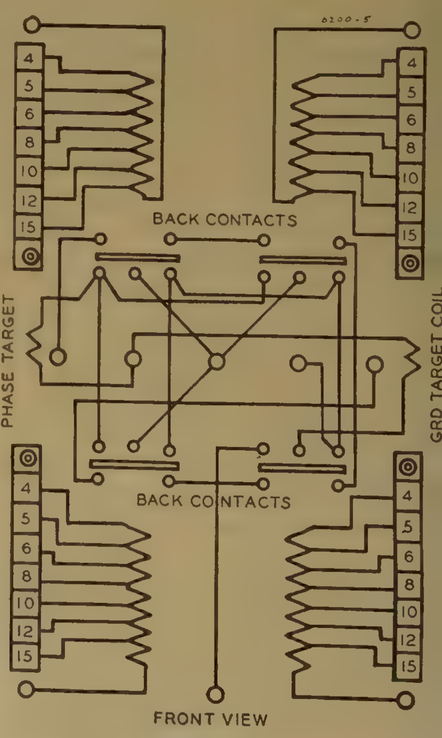

Figure 5. Internal connections of special fault detecting relay type *IT-12*

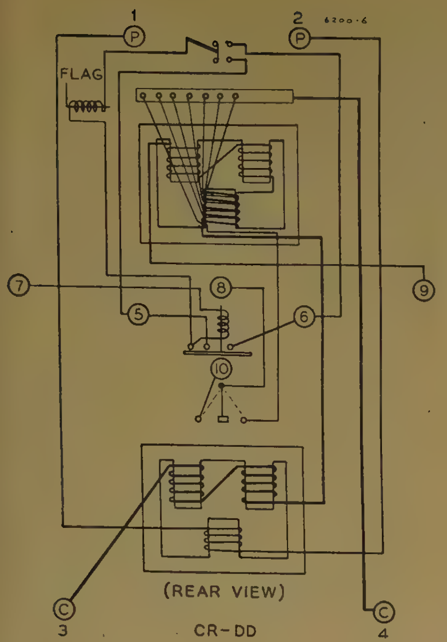

Figure 6. Internal wiring diagram of special ground relay type *CR-DD*

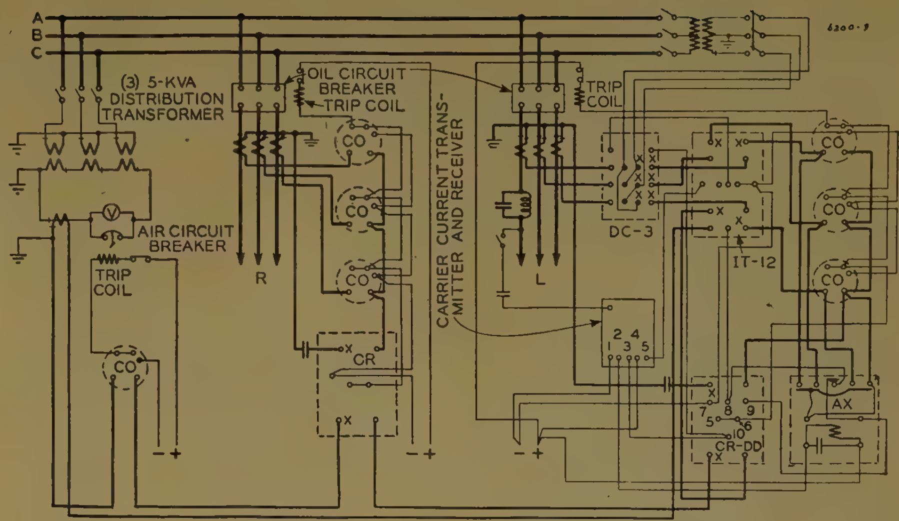

Figure 7. Receiving station relay system

Figure 7 is a diagram of the relay setup at receiving stations consisting of one radial feeder and one loop feeder. This setup is different in that there is no limiting resistor in the secondary delta circuit and a phase directional relay has been added for phase protection. The reason no phase directional relay is necessary at the source station is apparent in that any phase-to-phase short-circuit current is away from the station and not toward the station. The over-all operation of the system is simple and makes use of the fact that if fault current flows through a feeder toward a station bus, that feeder is unfaulted and should not relay. Therefore in Figures 1 and 2, ground current faults can flow toward any station in the loop feeders while phase-to-phase current can flow in the feeders only toward receiving stations. For example, should a ground fault occur on the radial feeder out of Katella, a source station, ground current will flow toward Katella on the two feeders from Orange and the two feeders from Anaheim. If a phase-to-phase occurred on this radial feeder, no short-circuit current would flow on these four feeders. If a ground fault occurred on a radial feeder out of Orange substation, ground current would flow in the two Orange circuits in the opposite direction from the foregoing case, while the current in the two Anaheim lines still would flow toward Katella. However, if a phase-to-phase fault occurred on this feeder, short-circuit current would flow toward Orange on its two circuits and none on the Anaheim lines.

This relay system will cause a directional relay to close contacts when fault current is flowing toward a station bus. If the fault is phase-to-phase it will cause the polyphase directional relay *DC-3* to close contacts, and if the fault is phase-to-ground it will cause the ground directional relay *CR-DD* to close contacts. If the fault current is of sufficient magnitude it will cause the fault detector relays *IT-12* to operate its contacts. The contacts of each of these two sets of relays are in series and energize the transmitting relay which causes the carrier set to transmit, sending a signal to the remote end of the distribution circuit to prevent the relays at that end from tripping. This is accomplished by having the receiver relay energize the lockout relay *AX*. The lockout relay short-circuits the operating coils of the over-current phase tripping relays and opens the directional control of the ground tripping relay, thus preventing the operation of the tripping relays. This is done also at the transmitting end by energizing the lockout relay directly from the transmitting relay *TR*. Figure 6 is an across-the-line diagram to show schematically the connections of these various relays.

Because the receiving substations are all unattended, it became necessary to devise a method of routine testing of the entire carrier relay circuits. The receiver relays are set to operate at two milliamperes current and the second receiver relay has a setting of five milliamperes. This second relay is used normally as a signal relay so that when reception is below five milliamperes, it will not function. Therefore, the scheme of testing must have the operator at the source substation transmit on one of the loop circuits. This will cause the signal receiver relay at the remote or receiving substation to operate, which in turn causes the transmitter of the other circuit at the receiving substation to transmit to the signal receiving relay at the source substation energizing a pilot light. This gives a positive check on carrier operation in that direction. By energizing the other transmitter at the source substation, a positive check is made for carrier transmission in the opposite direction. In Figure 1, if the transmitter is energized on the Altadena line at Eagle Rock, the receiver relays of the other two circuits should indicate proper transmis-

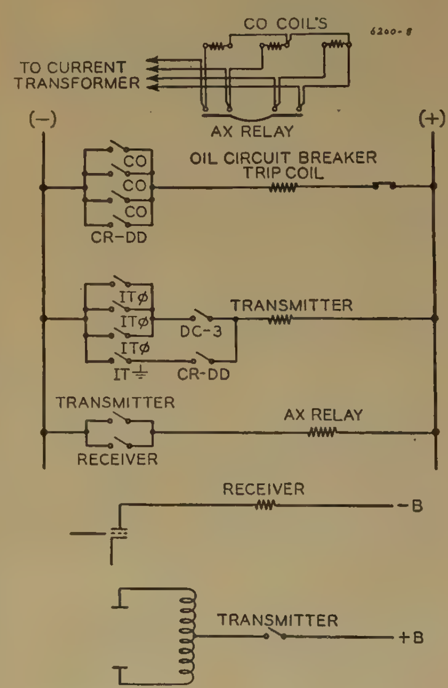

Figure 8. Across-line diagram of relay system

sion. This is true of the other two transmitters at Eagle Rock substation.

Operating Experience

The systems in Figures 1 and 2 have been in operation for six years, and very few difficulties have been uncovered. In one heavy severe rain and wind storm there were over 200 operations, none of which were unnecessary. These 200 operations consisted of many ground faults and phase-to-phase faults.

There has been trouble with the *CR-DD* relays in that they have operated with current in only one circuit. This should not occur and sometimes it is very difficult to adjust these relays to prevent such operation. The wave trap capacitors absorb moisture causing difficulty with the carrier operation. This seems to be the greatest cause of trouble in all carrier relay systems of the Edison Company.

With these years of operating experience it can be said that this scheme is a practical solution to the ground fault problem. A less complicated system may be developed in the future.

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Application Ratings of Indoor Power Circuit Breakers

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Synopsis: Increased cost involved in serving electric power to customers, and the aim to lower rates, have accelerated the need for most efficient utilization of existing generating and distribution equipment. The paper describes a method for loading power circuit breakers on a temperature basis for the purpose of taking advantage of their inherent current carrying ability. Several years of practical application of the basic theory has proved entirely satisfactory on various types of electric equipment and has, in many cases, established considerable savings on new investments. While these savings apply to a lesser degree to circuit breakers than to other major power equipment, their application in accordance with temperature, rather than name plate rating, has put the use of new and existing facilities on an equally sound engineering basis.

As a result of the increasing need for better utilization of high-voltage line capacities, the question arose some years ago as to what maximum safe overloads could be carried on existing line equipment during emergency conditions. It is the practice to consider the firm capacity of a load distribution center as the summation of the capacities (ratings) of all associated supply lines except one, assuming for discussion purposes that the characteristics of all lines under consideration are the same, so far as current carrying ability is concerned. Heretofore, the individual lines had been loaded to less than 100 per cent of their published ratings under normal operating conditions, and slightly overloaded only during emergency. This is not to imply that rational overloading of equipment had never been practiced before; in fact, certain transformers had been operated on a temperature basis for a number of years. However, the temperature limits used were, as a rule, those which were set up for the entire industry by the AIEE, or in a few isolated cases, determined by tests on the system. Certain power circuit breakers also had been included in these tests, but a practice of overloading equipment in general, on a true temperature basis, had not been firmly established. In the proposed setup, the optimum conditions would be to operate each line normally at 100 per cent of its true rating and apply correspondingly heavier overload during emergencies, even if a small percentage of the equipment life had to be sacrificed at each occurrence. The emergencies were estimated to last for relatively short periods of time in the majority of cases, that is, a relatively heavy overload might be expected for the first two hours, but after that time some relief would be provided by load transfer, and after four hours the load per line would, as a rule, be expected to be reduced to the normal capacity of the line.

Before such a setup could be adopted, however, various pieces of equipment, comprising a line from the source to the receiving end had to be investigated; thus also, power circuit breakers. In the early attempts to load equipment on a true temperature basis, very little information could be obtained from current literature on the effect of overload on circuit breakers, and the manufacturers were cautious in their statements regarding overload capacity because they were becoming aware of the fact that, among older types of circuit breakers, only those with practically new contacts would stay within the temperatures permitted by the AIEE Standards under full load conditions. The manufacturers were fearful that overload of 2- and 4-hour durations might tend to weaken the contact structure and, in the case of most types of oil circuit

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Table I

Circuit-Breaker Name Plate Ratings	Per Cent Service Factor	
	Copper Contacts	Silver Contacts
2.4-kv and 4-kv service		
600–1,200 amperes	75	100
2,000 amperes and above	65	90
All types of metalclad switchgear		100
13-kv service		
600–1,200 amperes	80	100
2,000 amperes and above	75	90
All types of metalclad switchgear		100

1. When circuit breakers, except for metalclad units, are installed in open air, these service factors are to be multiplied by a compartment factor of 1.1, and when installed in tight compartments with no ventilation, the service factors are to be multiplied by 0.9.

2. A service factor of 100 per cent is to be used for metalclad switchgear as long as conventional installations are adhered to, or in cases where a single metal-enclosed unit is installed in an existing concrete cell with restricted air flow on three sides. Where metal-enclosed units are installed, either as a single unit or in the conventional way, but where in both cases the head room is seriously restricted, an arbitrarily chosen service factor of 95 per cent should be used. The fact that the units have tightly fitting doors should be disregarded because, where in the manufacturer's judgment extra ventilation is required, they are designed with louvers, or openings are provided in places where they are most effective. In other words, the compartment factor is one.

3. The maximum temperature of copper directly connected to circuit-breaker terminals is to be limited to 70 degrees centigrade under normal conditions, and 90 degrees centigrade under emergency conditions lasting two days or less, except for metalclad switchgear where the limits should be 85 degrees centigrade under normal conditions and 105 degrees centigrade under similar emergency conditions.

4. The maximum temperature of copper directly connected to circuit-breaker terminals is to be limited to 80 degrees centigrade under emergency conditions lasting for more than two days, except for metalclad switchgear where the limit should be 95 degrees centigrade.

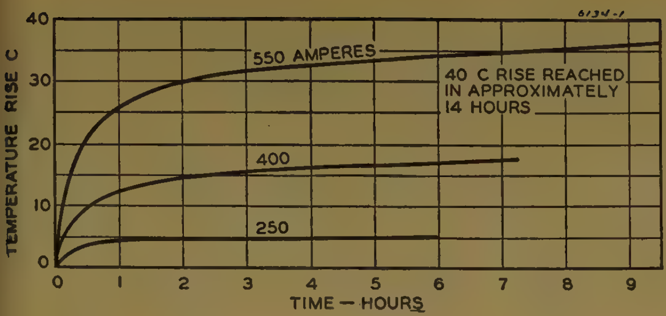

Figure 1. Temperature rise versus time curves for 500-ampere power circuit breaker

Main contacts in air

breakers, heat the oil to such an extent that the circuit breakers might catch fire if called upon to interrupt short-circuit current while in the overheated condition.

The only way really to determine what circuit breakers could do under overload conditions would be to make tests on typical units of various makes in the compartments in which they were operating.

Test, Evaluation, and Application

In view of the efforts made to provide for good housekeeping and proper maintenance of equipment, it was quite a surprise when tests revealed that the temperature rises on some of the circuit breakers substantially exceeded 30 degrees centigrade at full load (name plate rating) when tested in the "as found" condition. Some circuit breakers would not carry full load even when the main contacts were replaced with completely new assemblies obtained from stock. The loads at which the tested circuit breakers had 30 degrees centigrade temperature rise ranged from 45 per cent to 80 per cent of name plate rating in the "as found" conditions, and from 70 per cent to 100 per cent when reconditioned, depending on the type.

From the tests it could be stated quite conclusively that no oil circuit breaker with copper to copper contacts was able to carry current equal to its name plate rating without exceeding the standard 30 degrees centigrade temperature rise, except when in as good as new condition, and operating in open air or in a well ventilated compartment. Copper contacts, in particular, deteriorate after initial installation or subsequent overhauling, because of cumulative oxidization and the resulting increased temperatures.

It logically may be asked why this condition had not been discovered before tests were made. The answer is that circuit breaker ratings often exceeded cable ratings when single cable per line was used, that in the previous operating setup circuit breakers were called upon to carry full load only under rare emergency conditions, and under normal conditions the load currents were low enough so as not to make contact maintenance particularly excessive and thus did not arouse the suspicion of those in charge.

This is substantiated by the tests, and Figure 1 shows a typical example of the temperature rise of the main current-carrying contacts versus time, for three different load currents on a 15-kv 500-ampere oil circuit breaker of a type on which the main contacts are in air.

Deterioration of contacts due to cumulative oxidization, accidental burning during fault current interruption, gradual annealing and reduced contact pressure are all factors which tend to increase the temperature rise of a circuit breaker. The factor contributing the most, however, is probably oxidization. This means that in order to maintain rated temperature rise, the true rating of a circuit breaker ought to be decreased as a function of time after each reconditioning, but, since this is impractical, the ratings have been set up on an approximate basis, to secure rated temperature rise on the circuit breakers at the elapse of about one-half of the time between maintenance periods. This rating is referred to as the nominal rating.

Obviously, tests could not be conducted on all the various types of circuit breakers which are in service on the system, and they were therefore confined to eight units of the types most commonly used.

The method adopted for determining nominal ratings of circuit breakers, to be applied over a normal period between maintenance overhaulings, employs an assigned service factor and compartment factor dependent upon the circuit-breaker type, type of contact structure, and type of installation. The product of these factors and the name plate rating, will give a true rating for continuous load at which the circuit breaker is expected to have 30 degrees centigrade contact temperature rise above ambient temperature outside the breaker compartment, on the time basis explained above.

The list of service factors had to be chosen with considerable conservatism, particularly for metalclad switchgear, because operating experience with that type of equipment was considerably less at that time than it is today. In view of the fact that silver surfaced contacts are used in metalclad switchgear, and silver does not develop current-resisting oxides to the same degree as copper, experience has taught that there is less reason for conservatism with this equipment than with other types of switching equipment. The service factors now in use, excepting those of especially weak circuit breakers, are as listed below. These factors apply to circuit breakers in ventilated compartments, and are subject to modifications as stated in the notes supplementary to the list. Some of the factors recently have been increased slightly, and may be modified further if future experience so permits.

The purpose of the tests conducted on the eight circuit breakers mentioned was not only to determine the temperature versus time curves for these particular units, but also to give sufficient basic information so that the equations for such curves could be derived, and thereby a method established, by means of which the temperature rise for any combination of load and time could be calculated. The reader may sense that the entire analysis of this problem is rather involved. The practical application has been greatly simplified, however, by developing diagrams similar to Figure 2 for typical load cycles and for various types of circuit breakers.

Example: Normal* and typical emergency ratings are to be assigned to a 500-ampere 15-kv circuit breaker of the type used in the test, assuming that the ambient temperature is 40 degrees centigrade, the service

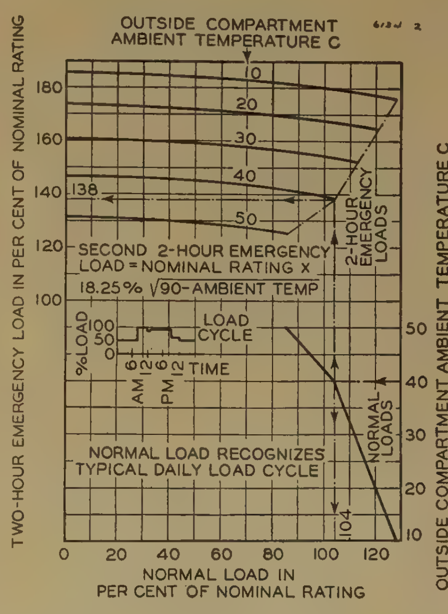

Figure 2. Permissible normal and emergency loads in per cent of nominal rating on indoor oil-poor and oilless power circuit breakers

Main contacts in air

Table II

Time, Hrs	Tested Temp Rise at 550 Amp	Calculated Temp Rise at 400 Amp	Tested Temp Rise at 400 Amp
0.5	21.2	11.25	9.5
2	29.8	15.8	14.5
5	33.5	17.7	16.7
7.5	35.0	18.5	17.5

factor is 0.8, and the compartment factor is 0.9; in other words, a circuit breaker with copper to copper contacts, mounted in a tight compartment and having a nominal rating of

$$500 \text{ amperes} \times 0.8 \times 0.9 = 360 \text{ amperes}$$

Following the guide on Figure 2 the results are

Normal rating $= 360 \times 1.04 = 375$ amperes

Two hour emergency rating $=$

$$360 \times 1.38 = 496 \text{ amperes}$$

Second 2 hour emergency rating is calculated from the formula

$$360 \times 0.1825\sqrt{90-40} = 466 \text{ amperes}$$

The basic theory on operation of electric equipment on a temperature basis as used in the Philadelphia Electric Company was developed by members[1-3] of this company's engineering department. The general equation has the form

$$\theta = A(1-\epsilon^{-at}) + B(1-\epsilon^{-bt}) \qquad (1)$$

where θ is the temperature rise, and A, B, a, and b are the equivalent of the factors which appear in the first two terms of an infinite series expressing the reaction of a thermal circuit to applied energy. These two first terms give a close approximation of the total series. Referring to the temperature rise curve in the 550-ampere test shown in Figure 1, and taking the slope of the curve at the times $t_2 = 3$, $t_3 = 6$, and $t_4 = 12$ hours, values designated $X_2 = 1.317$, $X_3 = 0.688$, and $X_4 = 0.425$ were obtained, which in turn give values for the parameters $A = 28$, $B = 12$, $a = 2.65$, and $b = 0.14$. Inserting these values in equation 1 gives the following numerical expression for the temperature rise versus time for this particular test:

$$\theta = 28(1-\epsilon^{-2.65t}) + 12(1-\epsilon^{-0.14t}) \qquad (2)$$

It may be noted from the curve that the final temperature rise in this test went to 40 degrees centigrade, and the preceding equation also satisfies this condition for high values of t. However, as it is desirable to limit the rise to AIEE standard value of 30 degrees centigrade, the equation for that condition becomes

$$\theta = 21(1-\epsilon^{-2.65t}) + 9(1-\epsilon^{-0.14t}) \qquad (3)$$

retaining the general shape of the curve while shifting its location. This step is justified by the tests because the temperature rise varies practically as the square of the current over the major portion of the curves, as will be seen from Table II and by referring to the temperature rises in the 400-ampere and 440-ampere tests. Following the same reasoning, the current which will cause a 30 degree centigrade temperature rise is therefore

$$I = 550\sqrt{\frac{30}{40}} = 475 \text{ amperes}$$

In the majority of cases when dealing with overload, the question is not so much one of finding out what temperature rise a circuit breaker will have when starting "cold," as it is to determine what total temperature it will reach at a given ambient temperature after having carried a certain amount of load prior to the application of overload. After all, it is the total temperature which sets the limit at which a circuit breaker can operate safely, and once this limit has been chosen, the corresponding current can be calculated.

If

X_t represents total contact temperature in degrees centigrade as a function of time, that is, temperature rise plus ambient temperature,

A represents ambient temperature in degrees centigrade,

t represents time in hours,

θ_{1o} and θ_{2o} represent the initial temperature rise in degrees centigrade of the component parts of total temperature rise before application of a load whose ratio to nominal rating is R,

θ_{1_f} and θ_{2_f} represent the final temperature rises in degrees centigrade of the component parts of total temperature rise at sustained load whose ratio to nominal rating is R,

then the equation for the total temperature can be written:

$$X_t = A + \theta_{2o} + \theta_{1o} + (\theta_{2_f} - \theta_{2o})(1-\epsilon^{-2.65t}) + (\theta_{1_f} - \theta_{1o})(1-\epsilon^{-0.14t}) \qquad (4)$$

where

$$\theta_{2_f} = 21R^2 \qquad (5)$$

$$\theta_{1_f} = 9R^2 \qquad (6)$$

and

$$R = \frac{\text{sustained load}}{\text{name plate current rating} \times \text{per cent service factor} \times \text{per cent compartment factor}} \qquad (7)$$

It may be noted that equation 4 becomes identical with equation 3 if the temperature rise is limited to 30 degrees centigrade, the initial temperature of the circuit breaker is the same as the ambient temperature, that is, $\theta_{1o} = 0$, $\theta_{2o} = 0$, and $R = 475/475 = 1$.

Since the tests, of which the results are show in Figure 1, were conducted on a circuit breaker whose main contacts are surrounded by air, equations 4 through 7 are believed to be accurate enough for practical application on any type of power circuit breaker whose main contact structure is in air.

To clarify the application of equation 4, let it be assumed that the 500-ampere test circuit breaker had carried 400 amperes continuously at 40 degrees centigrade ambient temperature, that the product of the service factor and compartment factor is 80 per cent, and that it is desired to find the total temperature which the circuit breaker will reach when carrying 200 amperes additional load for two hours. Graphically, the problem is shown in Figure 3 and the numerical solution is found as follows.

Because of the fact that the circuit breaker has been carrying 400 amperes continuously, prior to the application of the 600-ampere load, and its temperature, thereby, has reached a steady state, it is immaterial what the initial temperature rises were when the 400-ampere load was first applied. Equation 4 can therefore be written as follows for the steady state condition.

$$X_t = 40 + 0 + 0 + (\theta_{2_f} - 0)(1-0) + (\theta_{1_f} - 0)(1-0)$$

and inserting the values for θ_{2_f}, θ_{1_f} and R

$$X_t = 40 + 21\left(\frac{400}{500 \times 0.8}\right)^2 + 9\left(\frac{400}{500 \times 0.8}\right)^2$$

$$= 40 + 21 + 9 = 70 \text{ degrees centigrade}$$

These values, 21 and 9, become the initial temperature rises of the component parts so far as the following two hours are

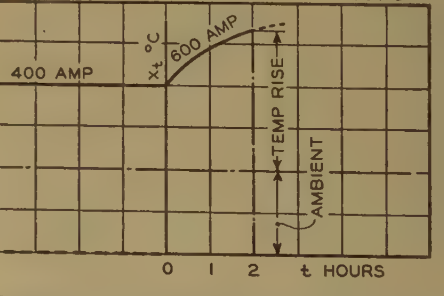

Figure 3. Total temperature of circuit breaker with 200-ampere load superimposed for two hours on 400-ampere normal load

* The normal rating is equivalent to the load which a circuit breaker can carry day in and day out without exceeding a total temperature of 70 degrees centigrade at 40 degrees centigrade ambient temperature, 60 degrees centigrade at 20 degrees centigrade ambient temperature, and 55 degrees centigrade at 10 degrees centigrade ambient temperature, recognizing the typical daily load cycle. Emergency loads equivalent to emergency ratings may be carried immediately following normal load.

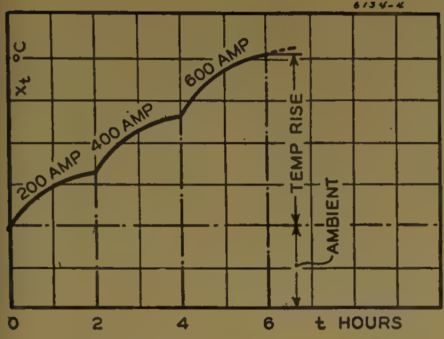

Figure 4. Total temperature of formerly idle circuit breaker after successive application of 200-ampere, 400-ampere, and 600-ampere load for two hours each

concerned, during which time the load is 600 amp. Therefore, equation 4 for that condition can be written:

$$X_t=40+21+9+\left\{21\left(\frac{600}{500\times0.8}\right)^2-21\right\}\times(1-\epsilon^{-5.3})+\left\{9\left(\frac{600}{500\times0.8}\right)^2-9\right\}\times(1-\epsilon^{-0.28})=98.9 \text{ degrees centigrade}$$

If, however, the conditions had been as shown on Figure 4, that is, the circuit breaker had been idle and had assumed the same temperature as the surrounding air before loads of 200 amperes, 400 amperes, and 600 amperes were applied at two hour intervals, the solution would be as follows:

First two hours

$$X_t=40+21\left(\frac{200}{500\times0.8}\right)^2(1-\epsilon^{-5.3})+9\left(\frac{200}{500\times0.8}\right)^2(1-\epsilon^{-0.28})=40+5.2+0.5$$

Second two hours

$$X_t=40+5.2+0.5+\left\{21\left(\frac{400}{500\times0.8}\right)^2-5.2\right\}\times(1-\epsilon^{-5.3})+\left\{9\left(\frac{400}{500\times0.8}\right)^2-0.5\right\}\times(1-\epsilon^{-0.28})=40+21+2.1$$

Third two hours

$$X_t=40+21+2.1+\left\{21\left(\frac{600}{500\times0.8}\right)^2-21\right\}\times(1-\epsilon^{-5.3})+\left\{9\left(\frac{600}{500\times0.8}\right)^2-2.1\right\}\times(1-\epsilon^{-0.28})=93.7 \text{ degrees centigrade}$$

It will be noted that by using equations 4 to 7, inclusive, and a step by step method of calculation, temperatures or ratings of oilless and oil-poor circuit breakers can be calculated for any type of load cycle, provided that in the first case, the load currents are known, and in the second case, the temperature limits are chosen.

For oil circuit breakers with all of the contact structure under oil, the same reasoning as above can be applied to the test results obtained on these types, and the total temperature of the contacts can be expressed by the equation

$$X_t=A+\theta_1+\theta_{2o}+(\theta_{2f}-\theta_{2o})(1-\epsilon^{-t/8}) \quad (8)$$

where

$$\theta_1=4.5R^2 \text{ for } t=\text{two hours or more} \quad (9)$$

$$\theta_1=7.5R^2 \text{ for } t=\text{less than two hours} \quad (10)$$

$$\theta_{2f}=25.5R^2 \quad (11)$$

and

X_t represents total contact temperature in degrees centigrade as a function of time, that is, temperature rise plus ambient temperature

A represents ambient temperature in degrees centigrade

t represents time in hours

θ_1 represents contact temperature rise above oil temperature in degrees centigrade

θ_{2f} represents final oil temperature rise in degrees centigrade under load whose ratio to nominal rating is R

θ_{2o} represents initial oil temperature rise in degrees centigrade before application of load whose ratio to nominal rating is R.

While some of the temperature limits now in use when rating circuit breakers have been stated in connection with the service and compartment factors, the selection of these and others needs to be justified.

The various tests were conducted on circuit breakers mounted in compartments which were either tightly closed with doors or partially ventilated, and the ambient temperature on which the temperature rises are based, was taken outside the compartment to make the basis consistent. For circuit breakers in tight compartments with no ventilation, it was found that the ambient temperature inside the compartment and thus the circuit breaker total temperature was approximately five degrees centigrade higher than for the same circuit breaker with the same load in a compartment with some ventilation. Based on this observation, the rating for circuit breakers in tight compartments should be approximately 0.9 times the rating of circuit breakers in ventilated compartments. Further, since the ambient temperature inside the compartment was approximately five degrees centigrade higher than outside the compartment, a circuit breaker installed on a structure without compartmentation, with little or no restrictions to the air flow, could carry approximately 1.1 times the load of the same circuit breaker when installed inside a ventilated compartment.

Table III. Average Ambient Temperature

Calculated (Deg C)	Use (Deg C)
17.5–22.5	20
22.5–27.5	25
27.5–32.5	30
32.5–37.5	35
37.5–42.5	40

Ambient Temperatures

Before any rating based on temperature can be assigned to a circuit breaker, it is necessary to arrive at an average ambient temperature at its location. In attended places, this is done by taking daily thermometer readings in the room every four hours during two seasons of the year, one season termed summer lasting from April through October and containing 214 days, the other termed winter lasting from November through March and containing 151 days. The daily maximum readings on the same thermometer are added over each of the two seasons and divided by 214 and 151 days respectively, thus arriving at a daily average ambient temperature for the summer, and likewise for the winter. If several thermometers are located in the same room the highest value calculated on the above basis is considered the average ambient temperature for that room. However, in any event, for the purpose of simplifying rating calculations, the average is rounded off as shown in Table III.

Continuous and Normal Load Conditions

Both terms cover day in and day out operation, except that the latter recognizes the typical daily load cycle for the individual problem.

The temperature rise of a circuit breaker is, as shown in Table II, nearly proportional to its I^2R loss. However, oxidization produces a steadily increasing contact resistance, particularly on copper to copper surfaces, and thus an increasing loss and temperature rise. If the total temperature is held constant and the time involved is not too long, this temperature will produce the same rate of oxidization, whether it is a result of high ambient temperature and low temperature rise, or vice versa. To clarify, compare the values in Table IV where it is assumed that under condition A, two identical circuit breakers are operating at different loads, at different ambient temperatures, but at the same total temperature when initially put in service. For a while the contacts are going to oxidize at the same rate, and let it be assumed that

after some time the contact resistance has increased 20 per cent. While initially there was no difference in total temperature, this no longer is true as the resistance increases with time, even though the ambient temperatures and the loads are held constant. This is shown under condition *B*.

Under conditon *B* it will be noted that circuit breaker 1 is four degrees centigrade hotter than circuit breaker 2. It will get progressively hotter, in fact, the heating is actually somewhat worse than indicated in the tabulation because the two circuit breakers will oxidize at different rates as soon as the smallest difference in total temperature appears.

From this analysis the conclusion can be drawn that, for circuit breakers operating on a continuous or normal basis, the allowable total temperature limits should be lowered with decreasing ambient temperature, although not necessarily in direct ratio. Stated in another way, the permissible temperature rise should be restricted with decreasing ambient temperatures. This restriction has been chosen as one-half of the amount between the actual ambient temperature and 40 degrees centigrade, as long as the ambient temperature is below the latter value. At 40 degrees centigrade ambient temperature and above, the total temperature has been limited to 70 degrees centigrade because, so far, no higher value appears justified under normal load conditions. It will be noted under condition *C* in Table IV that, in order not to exceed the permissible total temperature limits of circuit breaker 1 while operating continuously at loads exceeding its nominal rating, several remedies can be employed. These will be considered in the same order as tabulated under condition *C*.

Remedy 1. Reduce the load current. This may or may not be possible and would mean revenue loss unless it could be picked up on other less loaded circuits.

Remedy 2. Reduce contact resistance below original value. This means either to rebuild the circuit breaker for the next higher current rating, which generally is feasible only on circuit breakers rated below 1,200 amperes, or to replace the circuit breaker with one of higher rating. (It has been assumed that the new contact structure would have 20 per cent less resistance than the original in order to fit into this picture.)

Remedy 3. Reduce the ambient temperature by means of ventilation. As a result of ventilation, the compartment factor will change, which in turn will increase the nominal rating, and thus reduce the percentage overload (shown in parentheses, Table IV) for the same actual current. Thus, in this example, the contacts could, at least on a theoretical basis, be permitted to deteriorate until their resistance reached 150 per cent of the original before permissible temperature limits would be exceeded. Since the needs for overhauling, under the assumed conditions, would seem to be only a matter of time, the contacts might just as well be reconditioned and at the same time silver surfaced, which would permit carrying the same ampere load without resorting to drastic reduction of ambient temperature. (Remedy 4 refers to this alternate condition. Economy, of course, will dictate which of the two steps should be taken, because ventilation may in some cases be required for other equipment in the same room.)

Remedy 4. Recondition contacts. As already indicated in the preceding paragraph, reconditioning of contacts combined with silver surfacing proves as effective as, and is often much simpler than, any of the other remedies discussed, because it does not necessarily involve extra capital expenditure or loss of revenue. Since the silver surfacing will change the service factor, which in turn will increase the nominal rating, the percentage overload actually will be less for the same ampere load. (Figure in parentheses refers to this.) The same reasoning can be applied to circuit breaker 2 for silver surfaced contacts.

Silver surfaced contacts have much less cumulative oxidization than copper contacts because silver oxide is a relatively good conductor, while copper oxide is a relatively poor one. For this reason the service factors, determined on the basis of the tests, are higher for silver than for copper contacts. While all but the earliest tests actually were conducted on reconditioned circuit breakers, as good as new copper contact assemblies were used, and during the relatively short durations of the tests, these assemblies had practically the same contact resistance as silver surfaced contacts, although with time, they would deteriorate much more rapidly.

Short Time Emergency Conditions, Maximum Duration Two Days

As a result of the fact that contact oxidization under short time emergency overloads will be considerably less than that during long time overloads, 90 degrees centigrade total temperature limit at any ambient temperature has been chosen for emergency conditions lasting two days or less. This is on the safe side so far as effect of oxidization is concerned, but has been so chosen in order not to anneal the contacts, particularly where these are of laminated brush types. According to the circuit breaker manufacturers, the annealing of spring contacts is quite rapid at higher than 90 degrees centigrade, and oil will begin to smoke at about 100 degrees centigrade. Therefore, in order not to impair the interrupting ability of power circuit breakers while in over-heated condition, the 90 degree centigrade total temperature limit seems to be wisely chosen. This excludes metalclad switchgear. If oilless circuit breakers are used, 105 degrees centigrade is believed to be permissible. If oil circuit breakers are used in metalclad gears, special consideration will be given to the problem.

Since the 90 degree centigrade limit applies to emergencies of two days or less, the question may be raised as to what load a circuit breaker can carry for two days if the 90 degree centigrade limit is reached, for example, within the first two hours. Referring to equations 4 and 8, these take the form $X_t = 30R^2 + A$ under this condition. Solving for R, realizing that X_t already has reached 90 degrees centigrade and may continue at this value up to two days, the answer is

$$\text{Nominal rating} \times \sqrt{\frac{1}{30}(90-A)}$$

Long Time Emergency Conditions

Under long time emergency operation lasting more than two days, the total

Table IV

Condition	Circuit Breaker	Per Cent Resistance of Contacts	Continuous Load in Per Cent of Nominal Rating	Ambient Temp, C	Temp Rise, C	Total Temp, C	Remarks
A	1	100*	129	20	50	70	Exceeds permissible temp rise
	2	100*	100	40	30	70	Permissible total max temp
B	1	120†	129	20	60	80	Exceeds permissible total max temp
	2	120†	100	40	36	76	Exceeds permissible total max temp
C	1	120†	115	20	40	60	Limited by permissible total max temp
	1	80	129	20	40	60	Limited by permissible total max temp
	1	150†	129 (115)	0	50	50	Limited by permissible total max temp
	1	100‡	129 (115)	20	40	60	Limited by permissible total max temp
	2	100‡	100 (90)	40	30	70	Limited by permissible total max temp

* New or reconditioned copper contacts. † Oxidized copper contacts. ‡ Silver surfaced copper contacts.

temperature limit for power circuit breakers has been chosen at 80 degrees centigrade in order to avoid excessive contact oxidization. Similarly, metalclad switchgear with oilless circuit breakers is limited to 95 degrees centigrade. In metalclad gears, the contacts always are silver surfaced, and springs which furnish contact pressure are not depended on so far as current carrying ability is concerned. Again oil circuit breakers in metalclad gears are given special consideration.

Temperature Limits of External Connections

The total temperature limits for external copper directly connected to power circuit breakers have been so chosen that heat will not be carried into the circuit-breaker contacts. When apparent deviations occur at low ambient temperatures, they are not believed to be serious, as the circuit-breaker terminal structure has been found to act as a choke. The values are stated in Table I.

In the preceding discussion, except where applied to metalclad switchgear, the conventional 30 degrees centigrade rise over 40 degrees centigrade ambient temperature has been assumed in referring to circuit-breaker name plate ratings. Since the name plate ratings for metalclad switchgear are based on 45 degrees centigrade rise over 40 degrees centigrade ambient temperature, the temperature limits chosen under overload conditions for this type of equipment have been raised by 15 degrees centigrade, as compared with the limits applying to other types of switching equipment. Except for this difference, the same formulas and procedure are used on all types of indoor power circuit breakers.

Conclusion

The preceding discussion is believed to be comprehensive enough to enable those who have not yet tackled the problem of loading circuit breakers on a temperature basis to do so without difficulty. The basic problem is the same for all users, and while some of the details may vary, these details should be relatively easy to collect once the requirements are known.

In case of new circuit breakers, the calculated normal and emergency ratings may exceed name plate values by moderate amounts depending on the physical arrangement and the ambient temperature. On older types of circuit breakers, however, the name plate values will seldom be exceeded unless the units are equipped with silver surfaced contacts and ambient temperature, as well as the physical arrangements, are favorable.

The specific method described has been used on circuit breakers for several years on the Philadelphia Electric Company system and has given results which are believed to be entirely safe. The same general method also has been applied to other types of major power equipment and has resulted in considerable savings on capital investments. These savings have contributed to the maintenance of low rates in spite of the steadily rising costs involved in serving electric power to customers.

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Power System Development for Service to Hanford Plutonium Plant

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EARLY IN 1943 a number of engineers in the Bonneville Power Administration were informed of a new "mystery load" which was to be added to the BPA system. Later this load became known as the Hanford Engineer Works of the Manhattan District. We, of course, were warned that the new load was one of the war's top secrets. The product to be made, about which no one seemed to know anything, the location of the plant and its general arrangement, the amount of power to be taken by the plant, the location of the transmission lines which were to supply it, and the additional important fact that the power to be supplied had to have the utmost reliability were subjects which were not to be discussed unless absolutely necessary. It is our feeling that all these secrets were well-kept. For while BPA engineers had the responsibility of arranging their system to provide reliable service to this new load, there were times when we despaired of ever obtaining sufficient general technical information regarding the apparatus to be used by the Hanford Engineer Works to make possible the best co-ordinated system design.

At the time this new load first was considered on the BPA system slightly more than one year of war had passed. Materials and labor were critical, and the system had started serving many large war loads, including the reduction of aluminum and power for shipyards. Commitments also had been made to supply power to other large war factories, the plants for which were still under construction. A number of calculating board studies had been made previously of the transmission system including these future loads, and it was realized that the transmission system, heavily loaded as it would be, could not supply the type of reliable service which would be required by the new "mystery load." Accordingly, it was decided to determine by a series of a-c calculating board studies what system additions and changes would be necessary in order to obtain a reliable source of supply to the Hanford Engineer Works' load. These studies were started and included load-flow diagrams, transient stability studies, and relay studies, so that all possible contingencies could be taken care of.

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Location of the Hanford Engineer Works' Load on the BPA System

Geographically, the BPA's transmission system serves three major load areas—the Puget Sound, Portland, and Spokane areas. Figure 1 indicates the geographic location of 230-kv and 115-kv transmission lines, major substations, and generating stations as of June 1946.

The system peak load during the war was 1,427,000 kw. But in February 1943, when the Hanford Engineer Works first

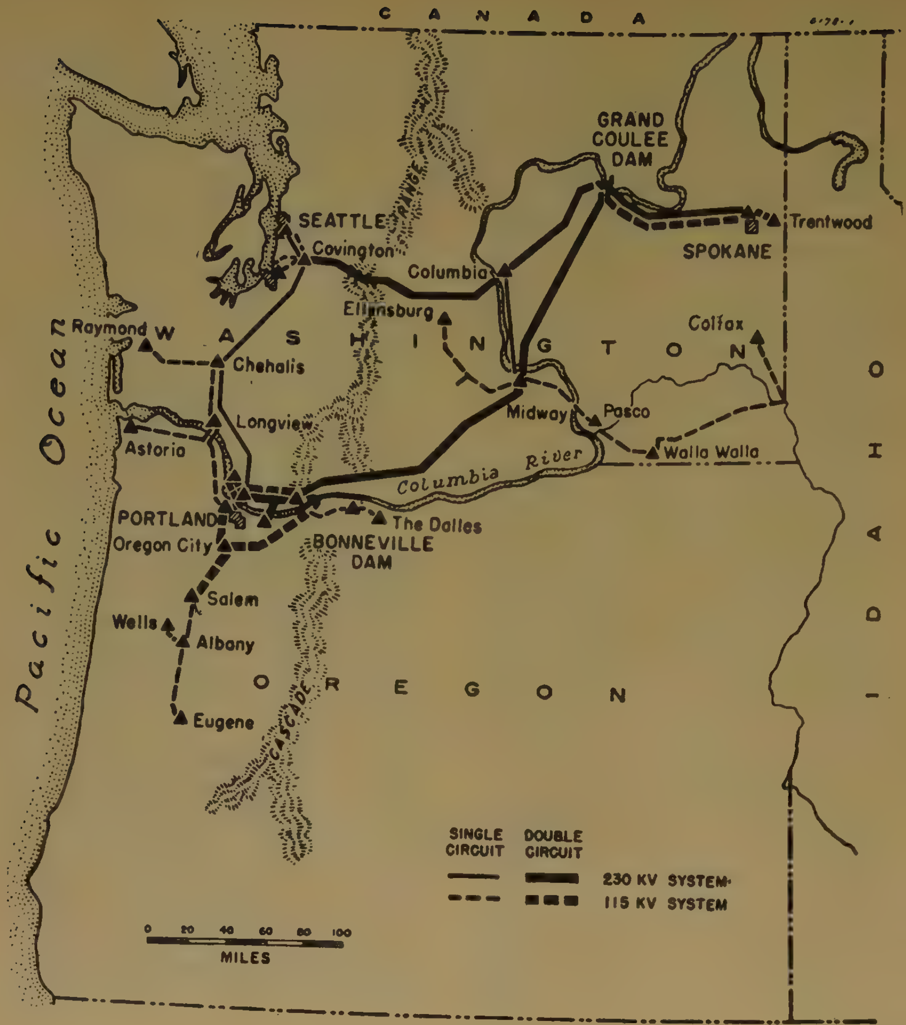

Figure 1. Bonneville Power Administration 230-kv and 115-kv transmission grid

was being considered as an additional load, the BPA system had not grown to its present size. A simplified schematic diagram of the BPA 230-kv transmission system as of February 1943 is indicated in Figure 2. This diagram shows a 230-kv loop with two large generating stations, Grand Coulee and Bonneville, and two large load centers, the Portland area and the Puget Sound area, connected to the loop. At this time there were three 108,000-kva generators installed at Grand Coulee and four 60,000-kva generators installed on the 230-kv system at Bonneville. But before the Hanford Engineer Works' load would be able to start operation, there would be eight Grand Coulee generators operating with two new 230-kv lines to Spokane and a large load at Spokane connected to the system.

The Grand Coulee generating station is constructed and operated by the United States Bureau of Reclamation; the Bonneville Power Plant is constructed and operated by the United States Army Engineers; and power from these two stations is transmitted to load centers over the transmission system of BPA. Two parallel 230-kv transmission lines connect the two generating stations through an intermediate switching station at Midway which is 100 miles from Grand Coulee and 134 miles from Bonneville. The Hanford Engineer Works' load was to be supplied from Midway, as this station had a source of supply either from Grand Coulee through two 230-kv lines or from Bonneville through two 230-kv lines.

The Problem

At that time the process to be used at the Hanford Engineer Works never had been tried fully so that knowledge of how it would work was very limited. In view of this lack of information and present-day knowledge of the destructiveness of atomic bombs, it is not surprising that extreme precautions were taken to maintain a source of supply of power to the load. We now know this power drove large pumps to supply cooling water to the uranium piles, and now can understand that the loss of power for even a short time might have brought about serious operational difficulties. Accordingly, it was decided, in addition to other less stringent reliability requirements, that the BPA system should be arranged to maintain power supply to Midway substation even for the remote probability of simultaneous or near simultaneous faults on both Coulee–Midway lines.

However, the following limitations were found with the 230-kv system as shown in Figure 2:

1. When all new loads and generation had been added to the system, loads in the Puget Sound and Portland areas would be very large in comparison with the transmission capacity, so that any double-line-to-ground fault on the Coulee–Covington line or on a Coulee–Midway line would result in the system being unstable. It would break apart, and in so doing would cause low voltage at Midway which was considered dangerous to the Hanford Engineer Works' process.

2. Power normally flows from Grand Coulee to Bonneville and the Portland area. After the system breakup occurred, with Grand Coulee separated from Midway, there was not sufficient generation at Bonneville to supply both the Portland area and the Midway loads so that the frequency and voltage at Midway would drop rapidly. Such drop in frequency also was considered dangerous to the Hanford Engineer Works' process. Similar difficulties would be encountered in case of a simultaneous fault on both Coulee–Midway lines.

3. At the beginning of the war, Grand Coulee had only one generator installed. Before August 1944 when the Hanford Engineer Works' load was to be energized, there would be eight generators installed with a maximum overload capacity of approximately 930,000 kw, and if all 230-kv lines and eight generators were to be connected to the same bus the short-circuit duty on 230-kv circuit breakers would approach 5,000,000 kva. Existing circuit breakers had an interrupting rating of 2.5 million kva while the largest available breaker, which was developed after the war started, had only a 3.5-million kva interrupting rating. Accordingly, the Grand Coulee bus had to be split up, and it was decided after considerable study that more reliable service could be supplied to the Hanford Engineer Works if one Coulee–Midway line was to be connected to each of the two new busses at Grand Coulee. This design of the Grand Coulee bus, which utilizes an interleaved arrangement of the transmission lines on two busses, has been covered in a previous paper.[1]

4. It already has been indicated how the system load, generation, and transmission lines had grown since the start of the war. Orders for high-speed carrier-current relaying had been placed for each end of all major 230-kv lines, but because of higher priorities of Army and Navy equipments, not one of some 22 sets ordered had been delivered. Without high speed clearance of faults at both ends of a faulted 230-kv line as provided by these relays, the system could

not be made stable without a greater number of reinforcements being required than otherwise would be the case. All such reinforcements, of course, had to be justified to the War Production Board because of shortages of critical material and labor.

Solution of the Problem

Extensive and intensive studies were conducted to determine the minimum system additions needed to obtain reliable service. Over 150 transient stability studies were made using several bussing arrangements at Grand Coulee and assuming faults at various locations on the system. The transmission system as shown in Figure 3, which finally was selected as satisfying the Hanford Engineer Works' load requirements for reliability as well as having sufficient capacity to supply other system loads, consists of the 230-kv loop with generation supplied at two points, while major loads are supplied from the loop at five other points. System characteristics were determined during transient stability swings following the clearance of a faulted line to be certain that relays would operate correctly when called upon to do so during these transient swings as well as during the faults. In this respect the war had forced transmission line loadings that were very high, and it was found that standard carrier type relays on some lines had to be modified before they would be completely satisfactory.

Because of the transmission loop arrangement, it was found that out-of-step blocking was undesirable, for instability might be initiated between any two major substations depending on the fault location, the initial system load division, or generation division, as well as circuit breaker and relay operating time. If such out-of-step operation did occur as a result of any cause, it was necessary to make certain that the system breakup would not produce low frequency or excessively low voltage at Midway except that low voltage would occur unavoidably for the few cycles (five to ten) between the time when a fault first occurred near the Midway bus and when it was cleared.

System Changes Needed

As a result of the foregoing and other studies, the following system changes and additions were made from February 1943 (Figure 2) to August 1945 (Figure 3). Important steps also are listed which are not shown easily on Figure 3.

1. Change in bussing arrangement at Grand Coulee.[a]

2. Addition of Coulee–Covington line number 2.

3. Arrangement for bus tie and bus sectionalizing breakers at Midway. Power to Hanford Engineer Works fed from both ends of the bus to prevent outage in case of a bus fault at the Midway substation.

4. Installation of carrier relays on all 230-kv transmission lines. Priority assistance to obtain delivery was needed most as most of the sets previously had been ordered.

5. Installation of a transfer tripping scheme at Coulee, Midway, and North Bonneville to provide an alternate source of supply from Bonneville generating station without drop in frequency or low voltage in the event of a double line outage of the two Coulee–Midway lines. (Until installation of Columbia–Midway 230-kv line.)

6. Installation of Columbia switching station (in Coulee–Covington lines) and Columbia–Midway 230-kv line.

7. Special relay studies.

(a). Resulting in modification of carrier relays.

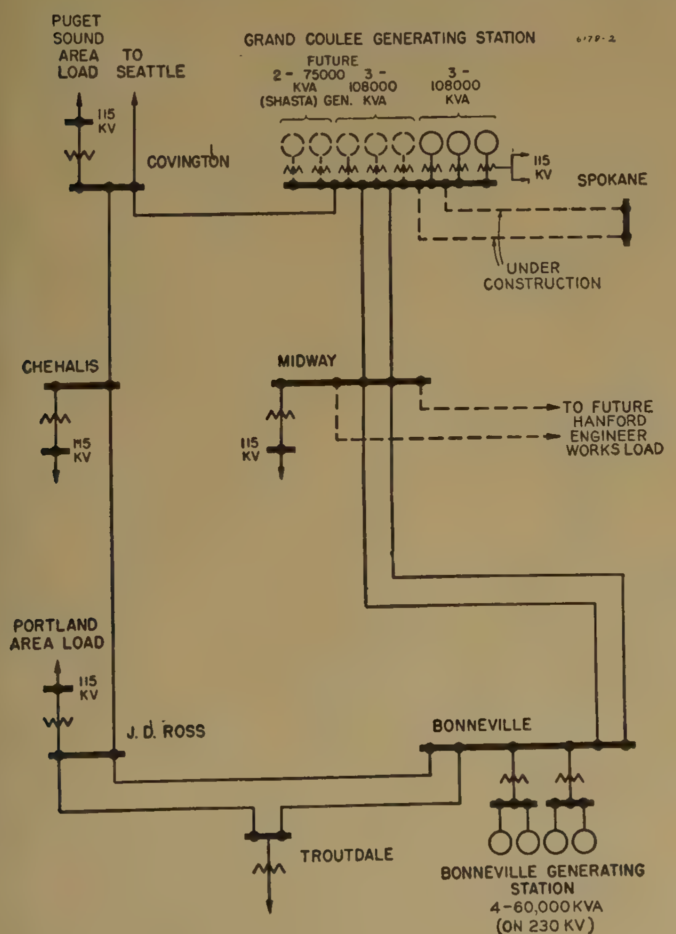

Figure 2. Schematic 1-line diagram of BPA 230-kv system in February 1943

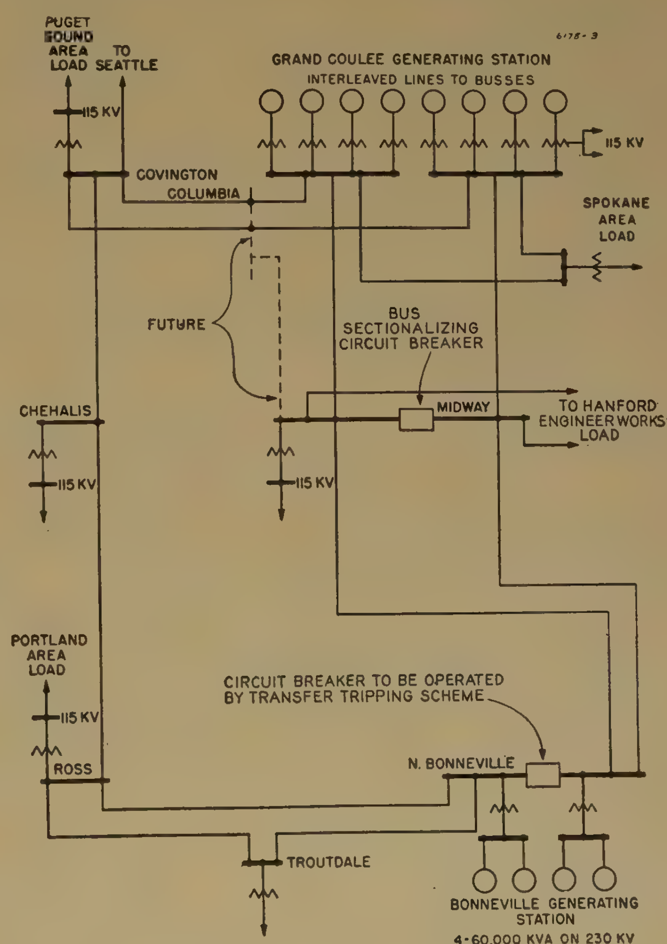

Figure 3. Schematic 1-line diagram of BPA 230-kv system in August 1944, approximate time when Hanford Engineer Works started operations

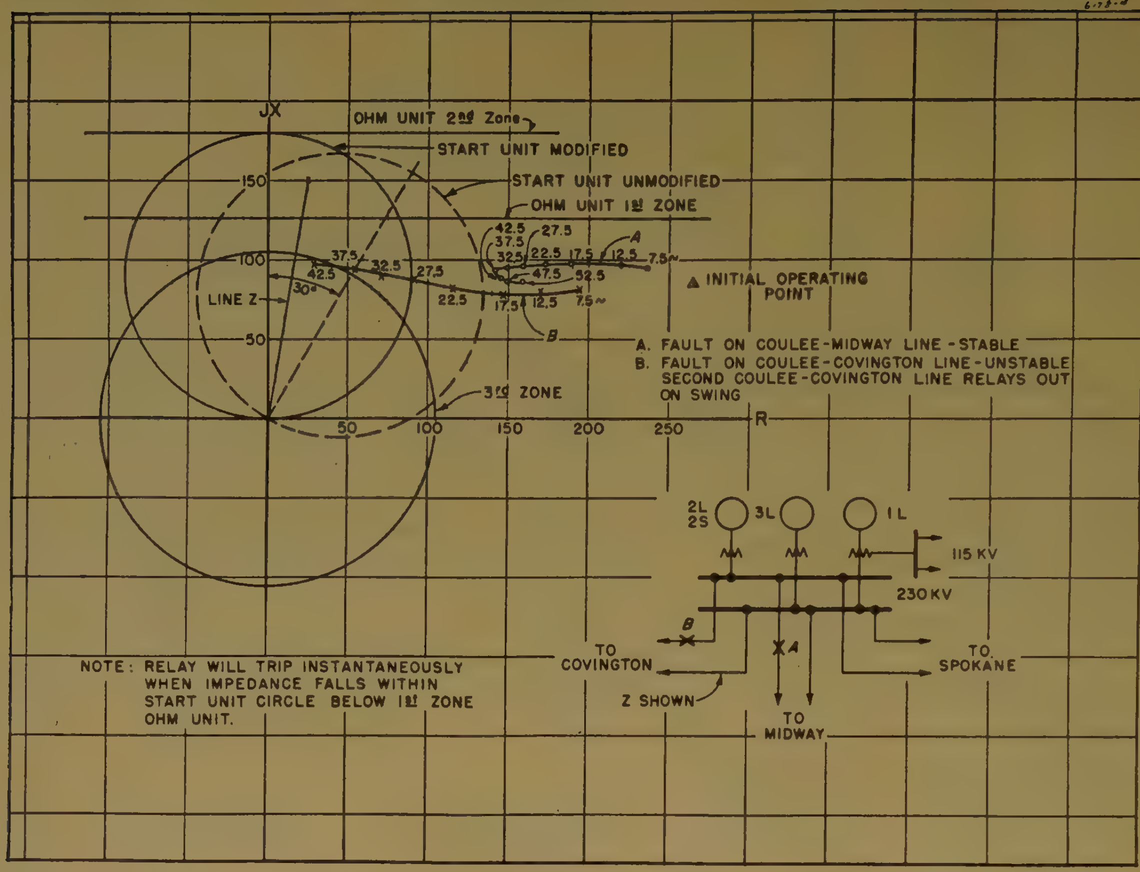

Figure 4. Impedances seen by Coulee end *GCX* an[d] *HDX* relays i[n] Coulee – Covingto[n] line during stabilit[y] swings

(*b*). Resulting in nonstandard relay settings.

(*c*). Special backup arrangements on relays.

8. System tests.

This list is rather imposing. Items 1, 2, and 4, and the Columbia switching station previously had been planned, but even though the load had grown enormously, lack of materials had stopped all construction and expansion except that so absolutely necessary that the War Production Board would grant priorities for the additions. A few of the foregoing additions and changes are of sufficient interest to be discussed more in detail.

Coulee–Covington Line

Repeated efforts had been made to obtain a second Coulee–Covington line long before anything was known about the Hanford Engineer Works' load; however, it had not been approved by the War Production Board because of scarcity of materials. When it became known that the Hanford Engineer Works' load would be connected, approval was expedited. However, before this line could be built and before the Hanford load was connected to the system, extremely heavy loads were carried on the first Coulee–Covington line. A peak value of 240,000-kw input at Grand Coulee was obtained. A board study previously had determined the static-stability limit of this line when connected to the system at Covington with two 108,000-kw generators connected at Grand Coulee at approximately 265,000-kw input. Hence the foregoing loading corresponded to 90 per cent of the static limit. For literally days at a time the line was loaded to 230,000-kw input (210,000-kw output at Covington). Over this period several faults occurred on the 115-kv system in the vicinity of Covington but several miles out from the bus. Nevertheless, the line was operating so near to the stability limit that the two generators connected to the line at Grand Coulee lost synchronism with the system for these faults, even though the fault normally would be considered remote from the generators. The circuit breaker at Grand Coulee would open for these conditions while the circuit breaker at Covington would remain closed. Further investigation showed that the relay characteristic needed to be modified at Grand Coulee to prevent tripping up to the maximum transient swing for which this portion of the system still would remain stable.

System Stability

With the Hanford load scheduled to be added to the Bonneville system, the procurement of carrier equipment, as well as all other materials and equipment needed to safeguard that load, became greatly facilitated. The original carrier relaying equipment consisted of standard distance relays (*HZ* and *GCX*), high speed carrier ground relays, and auxiliary carrier relays and equipment. These were to be arranged in well-known directional comparison schemes for high speed tripping of line faults by either first zone or carrier relays, and backup by other zones and by independent slower directional ground relaying. Ground current balance relays also were provided on parallel line sections. As mentioned previously, it was decided not to use relays which would block tripping on out-of-step conditions, but to do everything possible to avoid transient instability, and in those cases where instability was inevitable to permit the system to split where it would. Accordingly, transient stability studies were made of the system as shown in Figure 3, with the object in mind of determining the limit to which the system could swing and still remain stable, and the critical load and fault conditions which would produce such swings. During these studies the loci of apparent impedances seen by line relays were plotted on R-X diagrams and compared with the relay impedance characteristics.

It was learned that the 230-kv loop with generation at some points and loads

at other points would produce complicated loci of apparent impedances seen by line relays during these transient swings. An example of this is shown in Figure 4, on which is drawn the characteristics of a modified start unit and the unmodified start unit (shown dotted), for the *GCX* and *HDX* relays located at Grand Coulee in a Covington line. The loci of impedances, as seen by the relay during two transient swings, are drawn on this figure. Curve *A* corresponds to a fault on a Coulee–Midway line which after clearing was found to be stable, but the locus very closely approaches the unmodified circle before it turns away. The reversal of direction of this characteristic indicates that transient stability could be maintained by the system provided the relay does not trip. If the impedance had moved within the start unit circle, undesirable instantaneous tripping would be indicated. Curve *B* corresponds to a fault on one Coulee–Covington line, which after clearing was found to produce instability. It will be observed that the relays in the remaining Coulee–Covington line will trip this circuit when the apparent impedance has progressed until it is within the start unit circle. Numbers on each curve represent cycles elapsed after the fault first occurred; the first point was at 7.5 cycles because of calculating board technique. In all these studies, the fault was assumed cleared at both ends in five cycles through use of carrier relaying and 3-cycle breakers. As a result of these studies, the start unit was changed to the foregoing modified 90-degree characteristic instead of the standard 60-degree characteristic as shown dotted, since the 90-degree angle conforms more closely to the average impedance angle of BPA 230-kv lines ($81^1/_2$ degrees) than the 60-degree angle.

Another example of the transient impedance characteristics of the system, in this case compared with the *HZ* impedance relays at the Midway end of a Coulee–Midway line is shown in Figure 5. Curve *A* represents the impedance seen by the relay for a fault on the other Coulee–Midway line and is stable since the curve reverses direction. The curve actually progresses into third zone tripping (modified to be nondirectional) except that there is a time delay of 45 cycles before actual tripping can occur, and the system has moved out of danger before the necessary time can elapse. The directional element eliminates the possibility of the second zone tripping during this period.

Curve *B* shows corresponding values for a fault on a Coulee–Covington line which was unstable. It would appear that this relay also would trip; however, analysis of other relays on the system shows that the system breaks up between Coulee and Covington and between Midway and North Bonneville. Hence Hanford is left with power supply from Grand Coulee.

In Figure 5 it may be observed that the average reach of the second and third zones is rather short, being approximately 120 per cent of the line section. These were reduced in order to avoid tripping on transient swings from which the system can recover. These swings are shown in the figures as impedance loci which reverse direction, and it was to avoid having these swings get within the tripping areas for any appreciable time that the second and third zone areas were reduced.

In order to permit the reduced settings mentioned and at the same time obtain as much reach as possible for third zone backup time delay relays, these relays were made nondirectional. By this means the third zones of all breakers on a bus provide backup for any particular breaker, rather than depending upon remote breakers. They are better able to perform this function with the necessary reduced settings because they are one line section closer to the fault, and backup tripping tends to be localized to the affected bus section.

Special Backup Relaying

A time delay backup arrangement has been added in major stations which has definite advantages in meeting the re-

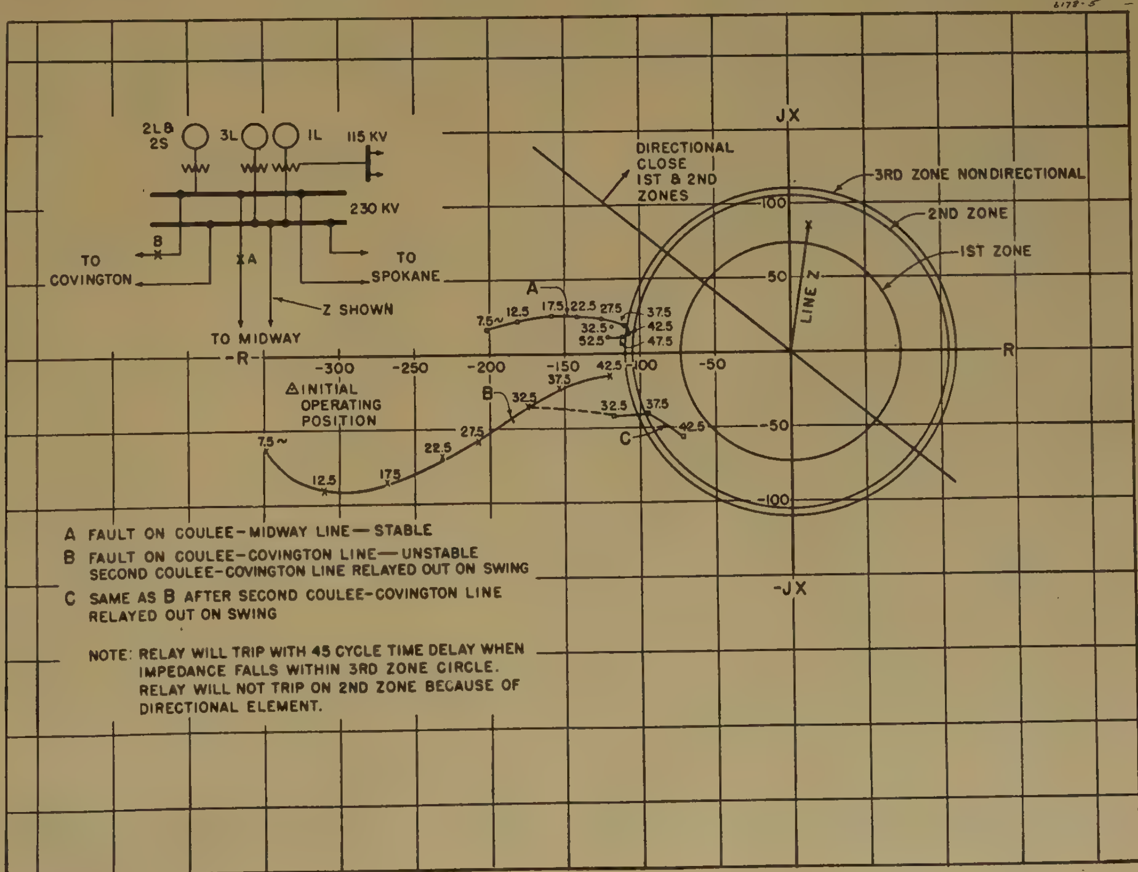

Figure 5. Impedances seen by Midway end *HZ* relay in Coulee–Midway line during stability swings

quirements of the system. This scheme, in the event that a breaker fails to open, operates the bus differential auxiliary relay associated with that breaker in 6 to 12 cycles after the breaker trip coil has been energized. (These are 3-cycle breakers.) This scheme tends to isolate trouble before second zone operations occur in other substations and is consistent with the attempt to localize backup tripping to the affected station as much as possible. Another important aspect of this backup feature, in cases where several sources can feed into a faulted line, is that it does provide backup where the normal type of overlapping directional impedance relays with time delay would fail to do so. The fact that the currents in the unfaulted lines are small percentages of the total fault current, because of multiple feeds, makes the impedance apparent to relays in the unfaulted lines often much larger than the settings of their backup zones. Since the normal backup scheme is not suited to the conditions, it was decided to use the special scheme described, the operation of which is initiated directly by the failure of the breaker to trip in the faulted line.

In order to provide more reliable service to the Hanford Engineer Works, at Midway substation only, the second zone distance trip circuits are connected to auxiliary relays arranged to trip any selected group of breakers. Since any second zone operation at Midway probably would be accompanied by system instability, this arrangement was made so that the system could be split to provide service to Midway from at least one line.

The third zone trip circuits of the Coulee and Bonneville lines at Midway are wired to trip one of the two bus differential auxiliary relays. The selection of the particular bus differential auxiliary to be tripped is made by directional relays on the bus sectionalizing breaker. Since both Midway–Bonneville lines are tripped by either bus differential auxiliary relay, this final backup arrangement of the Coulee lines. Similar backup for the Hanford lines is provided by instantaneous overcurrent relays which are operated by the sum of the two Hanford line currents. These relays cut out the third zone trip circuits but produce tripping through the directional relays and a timer, which is set for a longer time than the third zone delay of relays on the Hanford circuits.

Miscellaneous Relay Arrangements

A 1-cycle time delay has been added to all ground carrier trip circuits to decrease the possibility of false tripping because of gap flashover in carrier sets or sequential pole operation of oil circuit breakers.

Overcurrent fault detectors have been added in the first zone and carrier trip circuits of all the relays on the Hanford loop and many on the Bonneville system to prevent tripping caused by loss of potential.

Transfer Tripping Scheme

Certain network analyzer stability studies showed that the loss of both Coulee–Midway lines, before the installation of the Columbia–Midway line, during a critically heavy load period might result in system instability, that is, Coulee would run ahead of Bonneville and separate from the Portland area. As a result of this, Bonneville would be loaded heavily by the Portland area with resultant low voltage and frequency at Midway. To safeguard against this contingency, a bus sectionalizing breaker has been installed at North Bonneville (Figure 3) and a transfer tripping scheme arranged which can be used to isolate two Bonneville generators on the lines to Midway whenever both Coulee–Midway lines are opened.

This scheme is designed to operate as follows. If either line breaker is opened at Coulee, one of two audio tones is transmitted over the line carrier relaying channel to Midway. At Midway, tone relay contacts and pallet contacts on the Coulee breakers are arranged so that whenever both Coulee lines are opened at either end an auxiliary relay will initiate the transmission of two additional tones, both over the two carrier relaying channels to Bonneville. At Bonneville the audio components are filtered and brought to their respective tone relays. These relays have their contacts in series, and are also in series with a power directional relay, which is closed for power flow toward Midway, and with a 6-cycle time delay relay. This arrangement was made to prevent any possible misoperation of the transfer trip relays because of arcing on disconnects, which can cause the tone relays to operate.

To date, system loads have not approached the critical condition for which this scheme might have been necessary, and it actually is not in service.

Conclusions

Since the inception of the Hanford Engineer Works' load on the BPA system, no change on the 230-kv system or addition of any importance was contemplated without a consideration of its effect and the requirements of that load. The result has been the development of the afore-mentioned transmission and relaying system which was designed to supply continuous service to the Hanford Engineer Works, if necessary at the sacrifice of other loads on the system.

The extreme contingencies resulting from line faults, bus faults, or circuit breaker failure for which the system was designed have not occurred. The worst of this type which has occurred was a simultaneous outage caused by lightning of both Midway–Grand Coulee lines. This occurred twice in May 1946. However, the system was loaded lightly at the time, so that these faults, although severe, were not equal to the worst condition for which the system was designed. The Hanford operators noticed the voltage dip at the time, but they continued to operate normally, apparently without ill effects. The transfer tripping scheme was not in service because of a previous determination that it would be unnecessary under present system conditions, that is, with the Columbia–Midway line in service.

A more severe condition, not a result of a line fault, occurred in March 1946, when the excitation was lost accidently on five Grand Coulee generators, all that were operating at that time, and they pulled out of step from the rest of the system. The result was a drop in frequency to 94 per cent and in voltage to 72 per cent of the normal. Design information on the pumps and motors at Hanford indicates that even at this lowered voltage and frequency, sufficient power could have been supplied to keep the pumps operating. However, during this disturbance the pumping load was taken over automatically by a standby source of power within the Hanford area, thereby avoiding this critical test of apparatus.

Numerous other line outages, caused by a variety of line faults, have occurred without incident at Hanford other than momentary voltage fluctuations.

It is difficult to evaluate a transmission and relaying scheme when the goal is absolute continuity of service. Practical design considerations have been based on assumptions of very severe conditions, many of infrequent occurrence, through which service would have to be maintained, but there are always a number of imaginable disasters which unavoidably would result in loss of service. Generally speaking, it is believed that this scheme, as incorporated into the system, has met and probably will continue to meet the requirements of this load.

Reference

1. An Analysis to Determine the Optimum Bussing Arrangements and Transmission Capabilities at Grand Coulee, B. V. Hoard, G. W. Bills. AIEE Transactions, volume 63, 1944, pages 1259–64.

An Electronic Frequency Meter and Speed Regulator

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An electronic frequency meter and speed regulator has been developed to measure and hold the speed of a small steam turbine within plus or minus one-half of one per cent of any speed between 10,000 and 100,000 rpm.

This equipment fills a need for precise measuring and speed control apparatus required in the development and testing of very high speed machinery. In addition, the new circuits are of such basic nature that they readily can serve other applications for which no suitable commercial equipment is available. The frequency meter in its present form can be used as a general purpose laboratory instrument for the precise measurement of frequency or the accurate indication of small deviations from a given frequency. It gives an indication which is stable to about one or two rpm under normal line voltage variations, and its accuracy can be made to approach one-tenth of one per cent.

For use on machinery operating at speeds up to 100,000 rpm, the measuring instrument must not unbalance (and in some cases must not load) the moving part; therefore a photoelectric speed pickup is employed which requires no mechanical connection to the turbine shaft. The speed pickup, which can be placed several inches away from the turbine, merely "looks at" alternate black and polished areas on the surface of the turbine shaft and produces an electric signal whose frequency is some known multiple of the rotational frequency of the shaft.

Then in place of counting the revolutions per minute of the shaft we have the problem of measuring the frequency of an electric signal. The frequency meter used for this purpose must meet the following requirements:

1. Accurate within plus or minus one-half of one per cent or better.
2. Continuously adjustable over a frequency range of at least ten to one.
3. Capable of operating from an input signal having a very distorted wave form such as would be provided by a photoelectric or magnetic pickup.
4. Capable of providing an error signal proportional to deviation from any desired frequency to operate a servomechanism for regulating the turbine speed.

None of the commercial frequency meters are able to meet all of these requirements. A new frequency measuring circuit therefore has been devised which is essentially an electronic frequency meter of the recurrent capacitor discharge type but modified to operate as a "null-balance" circuit instead of a direct-reading device. By this simple modification the accuracy is increased from plus or minus two per cent to the order of one-tenth of one per cent and means are provided for obtaining a d-c or an a-c error voltage proportional to variation of shaft speed from the desired value. This voltage is used for indicating speed variations and for automatic speed regulation. It also operates a protective relay which will stop the machine in the event of an excessive speed deviation.

General Description

The desired speed is set up on a 3-dial decade speed selector having increments of 10,000, 1,000, and 100 rpm. The difference between selected speed and measured speed is called speed error. Its magnitude and algebraic sign are indicated on a large speed error meter, which has two ranges: plus or minus 250 rpm and plus or minus 1,000 rpm. A small speed meter, 0 to 100,000 rpm, is provided to show the approximate speed at which the turbine is running.

An electropneumatic servomechanism actuated by a speed error signal operates a steam control valve to regulate the steam pressure at the input to the turbine and hold the turbine speed at the desired value. The servomechanism is provided with a proportional air pressure "follow-up" and a manual reset control called the air pressure control dial. The steam control valve also is provided with proportional pressure follow-up so that the valve reproduces output pressure rather than valve position. In addition, the speed error signal contains a certain amount of error "anticipation" to minimize the response time of the system. By holding down a "bat handle" on the control panel the operator at any time can disconnect the speed error signal and assume manual control of the turbine speed, which he regulates by means of the air pressure control dial.

A block diagram of the complete speed control system is shown in Figure 1. The general scheme of operation is as follows: The turbine operator sets the speed selector dial to the number of revolutions per minute at which he desires the turbine to run. By means of the air pressure control dial he manually brings the turbine up to speed while watching the approximate turbine speed indicated by the speed meter. When the turbine speed comes within 4,000 or 1,000 rpm (operator's choice) of the desired preset value, the system switches into automatic control and the air pressure control dial becomes a fine speed adjustment. The operator trims the air pressure to make the speed error meter stabilize at zero. The frequency meter and servomechanism then hold the speed within plus or minus one-half of one per cent of the desired value. When a major load change is made, the operator retrims the air pressure control dial to return the speed error to zero, and the system regulates about the new control

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point. To measure a given speed at which the turbine is running, the operator adjusts the speed selector until the speed error meter read zero, and then reads the setting of the speed selector dial.

If for any reason the actual turbine speed goes either above or below the selected speed by more than 1,000 or 4,000 rpm (operator's choice), automatic protection circuits will shut off the steam supply and stop the turbine. In addition, the turbine automatically will be shut down by failure of any of the electric circuits. The speed error meter and automatic protection circuits function regardless of whether the system is in automatic or manual control.

The operator's control panel is shown in Figure 2. At the top is the frequency meter chassis with the speed selector in the center. On the left is the speed meter and an indicator light which shows when sufficient speed signal is present. On the right is the speed error meter, its range switch, and a light which indicates when the speed error is less than 1,000 or 4,000 rpm. Below the frequency meter is a panel containing steam and air pressure gauges. The next lower chassis contains the servo and automatic protection circuits, with the "bat handle" in the center and the air pressure control dial on the right. At the very bottom is the power supply chassis, with switch, fuses, and a monitor voltmeter.

At the present time, the turbine is being driven by high pressure air instead of steam, involving no changes, however, in the speed control equipment.

The Basic Frequency Measuring Circuit

The basic null-balance frequency measuring circuit is shown in Figure 3. It consists of a pair of push-pull connected triode switching tubes (V_1 and V_2) capacitor coupled to a diode bridge rectifier. The negative end of the bridge output resister R_B is grounded to the negative end of the common triode cathode resistor R_K and the negative side of battery B_1.

A square wave voltage, derived by the use of suitable shaping or trigger circuits from the signal whose frequency is to be measured is applied to the input terminals A and B in such a way that during one half-cycle the potential of terminal B is sufficiently negative to drive tube V_1 to cutoff. During the period of cutoff no voltage drop caused by tube current occurs in the resistor R_{L1}, and the capacitor C_1 will become charged to approximately the potential of battery B_1. The charging path is from ground, through battery B_1, resistor R_{L1}, capacitor C_1, diode tube V_4, rectifier output resistor R_B, and back to ground.

During the next half-cycle the signal applied to terminal B is within the operating range of tube V_1, and terminal A is sufficiently negative to drive tube V_2 to cutoff, so that tube V_1 will conduct an allow the potential of its anode to drop b an amount IR_{L1}, where I is the stead state plate current of the tube. The po tential across the capacitor C_1 will drop b the same amount IR_{L1}, the discharge pat being to ground through tube V_1 and re sistor R_K in parallel with resistor R_{L1} an thence back to the capacitor C_1 throug battery B_2 and diode V_5.

Since the electronic circuits are sym metrical, a similar charging and discharg ing of capacitor C_2 produces identical cur rent pulses at alternate half-cycle inter vals of the driving signal voltage, with th result that the charging current for th capacitors C_1 and C_2 flows through the re sistor R_B in the manner illustrated in Fig ure 3a, and the discharging current of th capacitors C_1 and C_2 flowing through th resistor R_K takes the form shown in Fig ure 3b. The resistor R_B, therefore, car ries only the transient charging current o capacitors C_1 and C_2, while the resistor R carries the steady current of the tubes V and V_2 in addition to part of the transien discharge current of the capacitors C_1 an C_2.

In order to utilize this circuit as a fre quency meter, it is only necessary to ad just discharge resistor R_K so that th average voltage developed across this re sistor equals the average voltage acros the rectifier output resistor R_B. Whe this has been done the voltmeter V wi indicate zero.

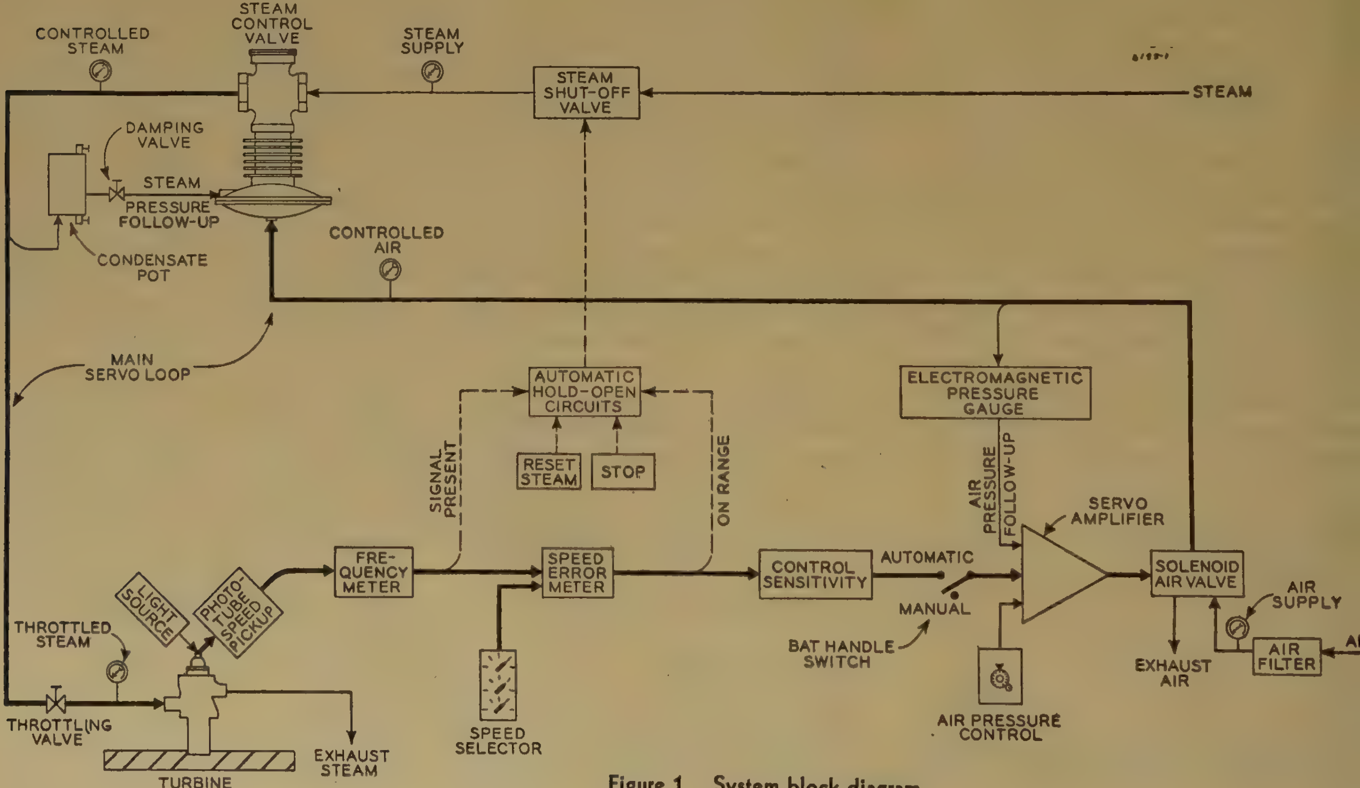


Figure 1. System block diagram

Figure 2. The control cabinet

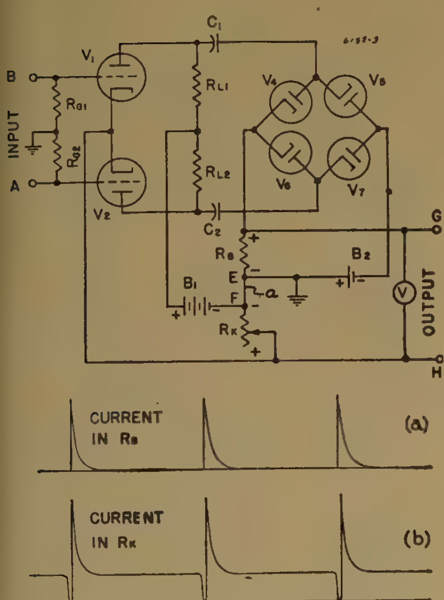


Figure 3. The basic null-balance frequency measuring circuit

If the frequency of the driving voltage applied to terminals A and B is raised, the heights and durations of the peaks in Figures 3a and 3b will not change, but will move closer together. The average current through resistor R_B is thus proportional to the frequency. However, the average current in resistor R_K is practically independent of frequency because the area of the discharge current peaks is only a fraction of the total area under the curve shown in Figure 3b. Consequently, the value of the resistor R_K required to balance the voltmeter V to zero will be very nearly proportional to the frequency of the signal applied to the input terminals A and B.

The equation of balance now will be derived. Since the circuit is symmetrical, either of the identical wire-wound resistors R_{L1} or R_{L2} can be represented by the single symbol R, and either of the high-stability mica capacitors C_1 and C_2 can be represented by the symbol C. R_p indicates the anode resistance, I the steady state anode current of either of the tubes V_1 and V_2, f the frequency of the signal voltage applied to the terminals A and B, R_B the value of the rectifier output resistor, and R_K the value of the discharge resistor required to balance the voltmeter V to zero. Each time a capacitor C is charged its voltage rises by an amount IR, and its charge rises by an amount CIR. The number of such charging pulses passing through R_B per second is $2f$. Thus the average voltage across R_B is

$$2fCIRR_B \tag{1}$$

Similarly, it can be shown that the average voltage across R_K is

$$R_K I+\left[\left(\frac{IR_K R^2}{R+R_p+R_K}\right)C\times 2f\right] \tag{2}$$

If R_K is adjusted until the voltmeter V reads zero, then these two average voltages are equal, and

$$2fCIRR_B=R_K I+\left[\left(\frac{IR_K R^2}{R+R_p+R_K}\right)C\times 2f\right] \tag{3}$$

Solving for f,

$$f=\frac{R_K}{2RC\left[R_B-\dfrac{RR_K}{R+R_p+R_K}\right]} \tag{4}$$

Since the expression $RR_K/(R+R_p+R_K)$ in the denominator varies with R_K from zero to a very small percentage of the resistance of R_B, the calibration curve of f as a function of R_K is very nearly a straight line, or

$$f\approx\frac{R_K}{2RCR_B} \tag{5}$$

If the circuit is to be used as a frequency meter, it is only necessary to vary the discharge resistor R_K until the voltmeter V reads zero, and then compute the input signal frequency by means of equation 4 or 5, or obtain the frequency value directly from the calibration curve previously prepared by comparison with a frequency standard.

Such a circuit can be used as a speed regulator control if discharge resistor R_K is adjusted so that the voltmeter V reads zero when a pickup device on the shaft rotating at the desired speed is supplying substantially square wave voltage impulses via a suitable trigger circuit to the signal input terminals A and B. Variations in speed of the shaft then will produce a pulsating d-c error voltage at the output terminals G and H which can be smoothed by a suitable electric filter and impressed upon a servo amplifier for speed control purposes.

Equation 5 shows that the only components which materially affect the calibration are the capacitor C and the resistors R_K, R_B, and R (R_{L1} and R_{L2} in Figure 3). If these components are stable and have low temperature-coefficients, the performance of the circuit as a frequency meter or as a speed regulator will be accurate and stable. The stability of the circuit also is increased by the fact that the voltages across resistors R_B and R_K are both proportional to the steady state anode current I so that variations caused by tube characteristics and battery voltages cancel out. The battery B_2 reduces to a low value the current which initial velocities of electrons in the diodes would cause to flow through the diodes and R_B. Changes in the contact potentials of the tubes resulting from changes

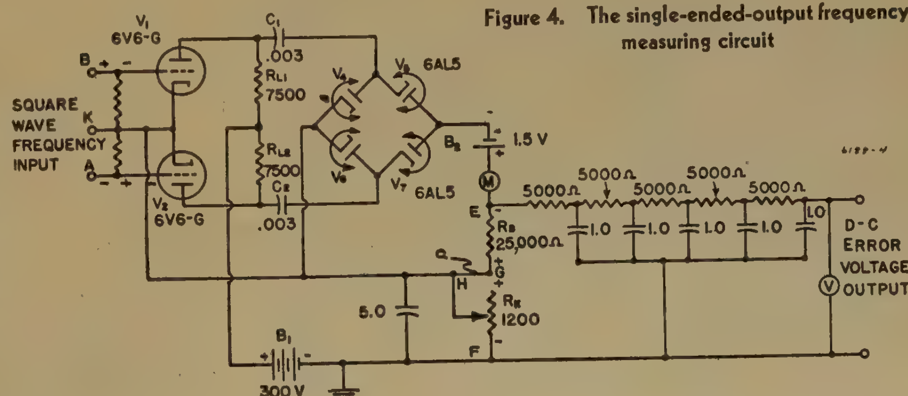


Figure 4. The single-ended-output frequency measuring circuit

in heater voltage are small compared with the charging potential of the battery B_1. Equation 4 shows that moderate variations in the anode resistance R_p of the tubes V_1 and V_2 have only a minor effect on the frequency calibration.

The Single-Ended Frequency Measuring Circuit

The circuit shown in Figure 3 and discussed thus far utilizes a double-ended input signal balanced to ground and produces an error voltage such that neither output terminal is grounded. A single-ended input can be utilized, if desired, by eliminating half of the symmetrical capacitor charging and discharging circuit, namely, V_2, R_{L2}, C_2, V_6, V_7, and R_{G2}. The output circuit can be modified to produce an error voltage with one output terminal grounded, by interchanging the wire *a* (Figure 3) with the voltmeter *V*, so that terminals *G* and *H* are joined together when the voltmeter is connected between terminals *E* and *F*.

The turbine speed control system described in this paper employs the frequency measuring circuit arranged for a double-ended input and grounded output, as shown in Figure 4. In this modified circuit the voltmeter *V* is connected between terminals *E* and *F*, and the terminals *G* and *H* are connected together. This joins the two measuring resistors R_B and R_K at their positive voltage ends rather than at their negative ends (as was done in Figure 3). In the circuit of Figure 4 at the bridge output, resistor R_B will conduct capacitor discharging current instead of charging current, but the circuit performance will be substantially the same as in the basic circuit which is shown in Figure 3.

Also in Figure 4, a milliammeter *M* has been inserted to measure the current in R_B. As this current will be proportional to the input frequency but subject to errors caused by variations in line voltage and tube characteristics, the meter *M* is used only for an approximate indication of the input frequency or speed. But the value of R_K required to balance the voltmeter *V* to zero at a given input frequency will be independent of variations in line voltage and tube characteristics to an accuracy approaching one-tenth of one per cent. The setting of R_K which gives zero output error voltage is used as a precise measure of the input frequency. The capacitor connected across the cathode resistor R_K, and the five *RC* sections connected to output terminals *E* and *F* serve to filter the pulsating output voltage into a steady d-c error signal.

Figure 5. Block diagram of the complete frequency meter

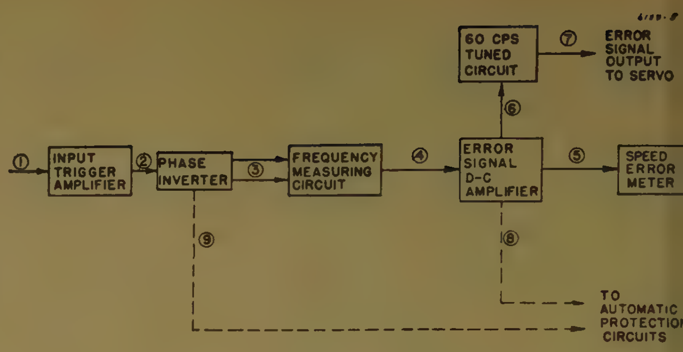
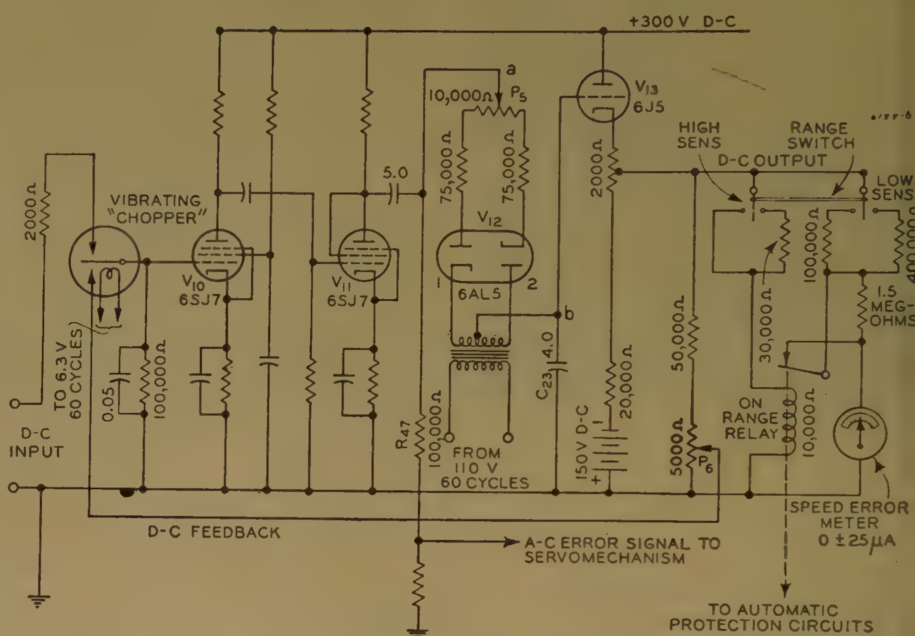


Figure 6 (below). The error signal d-c amplifier circuit



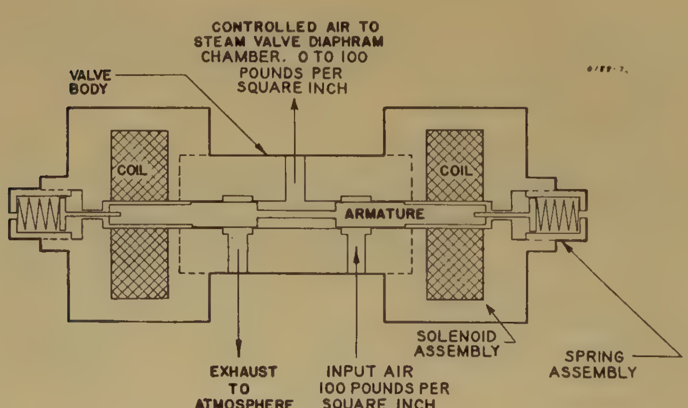


Figure 7. Schematic diagram of the solenoid air valve

The Complete Frequency Meter

A block diagram of the complete frequency meter is shown in Figure 5. An input trigger amplifier converts the low level input frequency signal (this can be of sine, square, or distorted wave form) into high level single-ended square wave which is applied to a phase inverter. Output from the phase inverter is double-ended square wave which drives the two switching tubes in the frequency measuring circuit.

Full scale d-c error signal from the frequency measuring circuit is about 25 volts for 100,000-rpm speed error or 0.25 millivolts per revolution per minute deviation from the selected speed. In order to achieve a measuring and control accuracy of plus or minus one-half of one per cent if any speed between 10,000 and 100,000 rpm the circuit should be sufficiently sensitive to indicate changes of at least 0.1 to 0.05 per cent. For 0.1 per cent speed deviation the d-c output error voltage is 2.5 millivolts at 10,000 rpm and

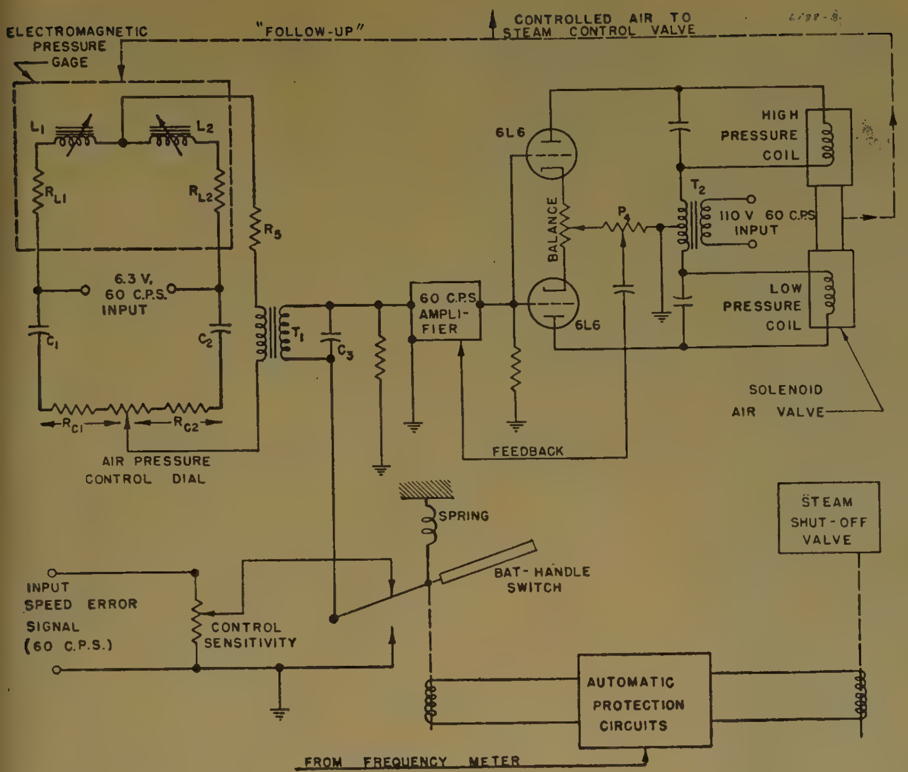


Figure 8. The servo circuit

25 millivolts at 100,000 rpm, which is not sufficient signal to operate a high resistance d-c voltmeter directly. Therefore the output voltmeter V shown in Figures 3 and 4 should be a vacuum tube voltmeter, which generally employs an amplifier to drive a panel type d-c microammeter or milliammeter.

The Error Signal D-C Amplifier

Because of the instabilities associated with conventional direct-coupled amplifiers a special stabilized, sensitive, discriminating d-c amplifier was devised for use in this frequency meter. This amplifier utilizes the established technique of "chopping" the d-c signal into square wave alternating current and passing it through a more easily stabilized a-c amplifier; but in addition it employs a familiar phase-discriminating rectifier to change the alternating current back to high level direct current, and incorporates d-c negative feedback to produce a high degree of over-all stability.

The circuit of the error signal d-c amplifier is shown in Figure 6.

The input d-c "chopper" is effectively a single-pole double-throw mechanical switch which opens and closes continuously at the rate of 60 cycles per second. It consists of a vibrating reed contactor driven between two stationary contacts by means of a 6.3-volt 60-cycle-per-second exciter coil. As the reed oscillates between the two fixed contacts it assumes the potential of the upper contact for approximately one-half of the 1/60-second cycle and the potential of the lower contact during the other half of the cycle. Therefore, when a direct voltage is applied between the two fixed contacts it will be converted to a 60-cycle-per-second square wave voltage across the vibrator output impedance connected between the reed and the lower fixed contact. The output alternating voltage will be in phase with the 110-volt 60-cycle-per-second line when the applied direct voltage is of a given polarity (say positive) and 180 degrees out of phase with the line when the input direct current is of the opposite polarity (negative).

Consider for the present that the lower fixed contact is grounded. Then the amplitude of the 60-cycle-per-second square wave from the vibrator will be approximately equal to the d-c error voltage and proportional to speed error. The phase will reverse when the speed error goes through zero. Connected across the 100,000-ohm vibrator output resistor is a 0.05-microfarad capacitor which serves to by-pass the high frequency transient peaks which occur during the "make" and "break" switching intervals.

The low level a-c error voltage is amplified in the conventional 6SJ7 stages V_{10} and V_{11}, and then impressed upon the phase-discriminating half-wave rectifier V_{12}, which operates as follows: Via the injection transformer T_1, the 110 volt a-c line voltage switches the two diodes of V_{12} on and off at the rate of 60 cycles per second and in synchronism with the 60-cycle-per-second vibrating reed in the d-c "chopper" at the input to the error signal amplifier.

The 60-cycle-per-second speed error voltage applied between point a and ground cannot reach point b when the diodes are cutoff; but during the half-cycle of line voltage that the diodes conduct there is a path of finite resistance from a to b allowing pulsating direct cur-

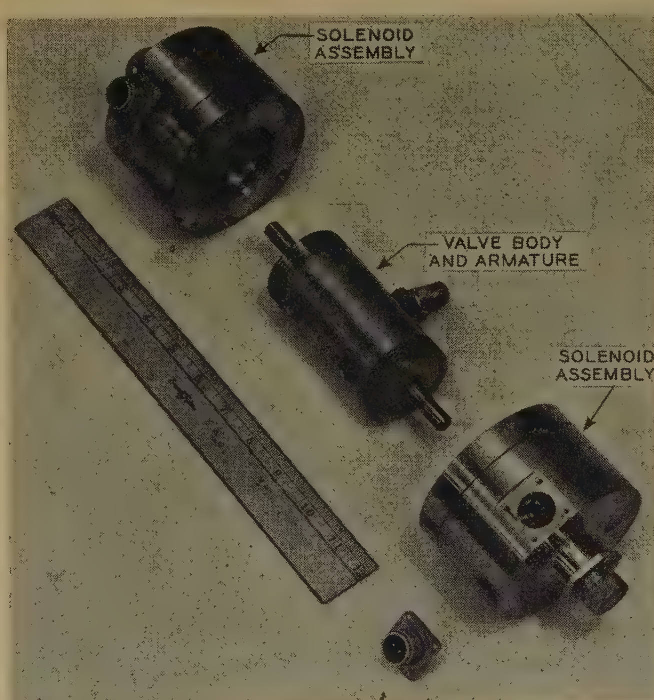


Figure 9. The solenoid air valve partly disassembled

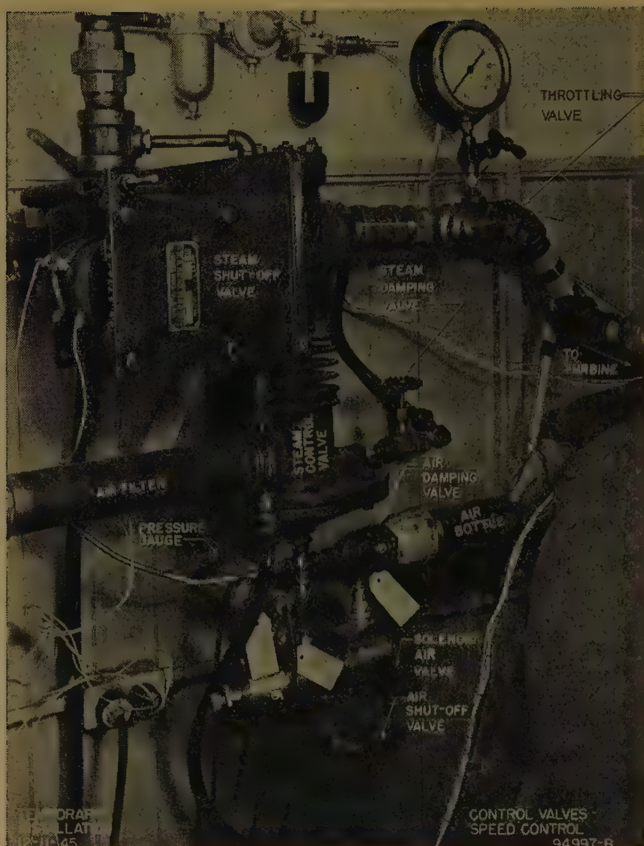


Figure 10. Temporary control valve installation with turbine operated by high pressure air instead of steam

rent to be fed to capacitor C_{23} through two parallel paths, each consisting of one-half of P_5, one 75K resistor, one diode, and one-half of the secondary winding of T_1. (The diodes are balanced so that the switching voltage from T_1 causes no potential difference between points a and b.) Because it comes from the d-c chopper, the a-c signal applied to point a is either in phase or 180 degrees out of phase with the 110-volt 60-cycle-per-second line. If a becomes positive when the diode plate 2 becomes positive the pulsating direct current will flow from a to b and through C_{23} to ground, producing across C_{23} an output voltage such that point b is positive with respect to ground. Similarly if point a is driven negative when diode plate 2 is negative, then the pulsating direct current will flow from ground through C_{23} and from b to a, producing across C_{23} an output voltage such that point b is negative with respect to ground. Capacitor C_{23} smooths the pulsating direct current to give a steady output voltage of the same polarity but greater magnitude than the d-c speed error signal applied to the d-c chopper at the input to the amplifier. The output high level d-c error voltage is impressed upon the cathode follower V_{18} which drives the zero-centered speed error meter to the right or left of zero to indicate positive or negative speed deviations.

By means of a voltage divider connected across the cathode follower output terminals, a portion of the d-c output voltage is fed back to the lower fixed contact of the input "chopper." The circuit is phased, by means of T_1, so that the potential at h is of the same polarity as the input potential at the upper fixed contact

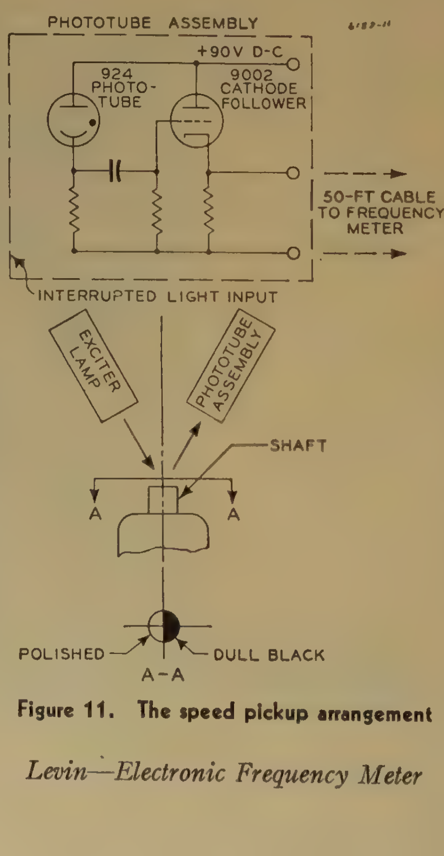


Figure 11. The speed pickup arrangement

of the vibrator. Therefore, as the vi brating reed assumes during one half cycle of 60-cycle-per-second line voltag the input potential of the upper fixed con tact and during the other half-cycle th feedback potential of the lower fixed con tact, the magnitude of the alternatin voltage output from the vibrator is re duced by the d-c feed-back voltage. Thi reduces the over-all gain of the error signa amplifier but serves to stabilize the ampli fier output against changes in supply volt ages and tube characteristics. Th amount of feedback, and consequently th gain of the amplifier, is adjusted by mean of P_6.

Also connected across the d-c output i the on-range relay which actuates th automatic protection circuits when th speed error exceeds 1,000 or 4,000 rpm depending upon the setting of the rang switch. When energized it also inserts a 1.5-megohm protective resistor in serie with the speed error meter.

The Input Trigger Amplifier

There are various shaping amplifier which can be used to operate the frequency measuring circuit. The one selected for this frequency meter is a new reliable "go–no go" circuit. That is, when the input speed signal amplitude exceeds a critical minimum voltage of about 0.15 volt (at any frequency from zero to at least 5 kc), the amplifier will trigger and provide high level square wave output of constant amplitude and one-half the input frequency. If the input signal amplitude falls below the critical value, the circuit will give no output at all. This feature is utilized by the automatic protection circuits to stop the turbine and summon the operator in the event of erratic or insufficient input signal from the photoelectric speed pickup.

The circuit diagram is shown in Figure 13. The input tube V_0 is an amplifier to feed the required 10,000-ohm input to the first trigger tube V_1. With the plate, cathode, and grid resistance values as shown, V_1 can be set up to have two stable operating points. The bias and feed-back resistors are adjusted until the tube is operating in one equilibrium condition and almost at the point where it will change to the second equilibrium condition. Then a small input signal will trigger the tube from one operating condition to the other. Output can be taken between ground and either the plate or the number 2 grid.

Although the tube will be triggered by either a rising or decreasing wave front as soon as the amplitude crosses a given

critical value, better square wave results when the triggering impulse is a sharp pulse, in which case a small capacitor connected between the plate and the number 1 grid adds to the reliability of operation.

The output from the number 2 grid of V_1 has squared sides but not a perfectly flat top and bottom. This signal is differentiated in the output R-C branch (25 micromicrofarads, 10,000 ohms) to obtain sharp pulses for tripping a second trigger tube V_3.

The differentiated output appears at the number 1 grid of V_2 as a series of alternate positive and negative pulses. But the time interval between a given positive pulse and the adjacent negative pulse might be different in length from the interval between the negative pulse and the following positive pulse if the speed input signal has unequal positive and negative half cycles such as might be provided by a photoelectric speed pickup. Therefore the blanking amplifier V_2 removes alternate pulses (by successive positive and negative blanking), producing at its second cathode a new series of equidistant pulses that are positive only. These are differentiated in the output R-C branch (10 micromicrofarads, 500,000 ohms) to obtain very sharp double-ended pulses which are used to fire the second trigger tube V_3. A given pulse triggers the tube in one direction and the following pulse trips it in the opposite direction. The resultant output from V_3 is a good square wave voltage having equal positive and negative half cycles. Its frequency is exactly half that of the input speed signal. This single ended square wave is fed to a 6N7-G phase inverter which produces double-ended square wave to drive the two switching tubes V_1 and V_2 in the frequency measuring circuit.

The Servomechanism

As shown in Figure 1, the diaphragm-actuated steam control valve is operated by an electropneumatic servomechanism, which receives 60-cycle-per-second speed error signal from the frequency meter unit. To supply regulated air to the steam control valve, at 0 to 100 pounds per square inch, the servomechanism utilizes a solenoid air valve shown schematically in Figure 7.

The air valve contains a balanced armature that slides in a cylinder which is provided with an annular metering port on each end and a central "controlled pressure" outlet connected to the diaphragm chamber of the steam control valve. At

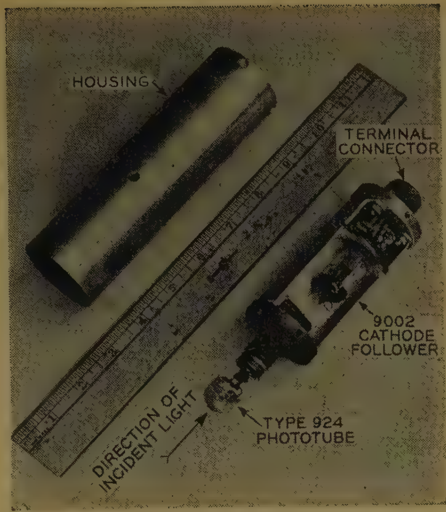


Figure 12. The phototube assembly

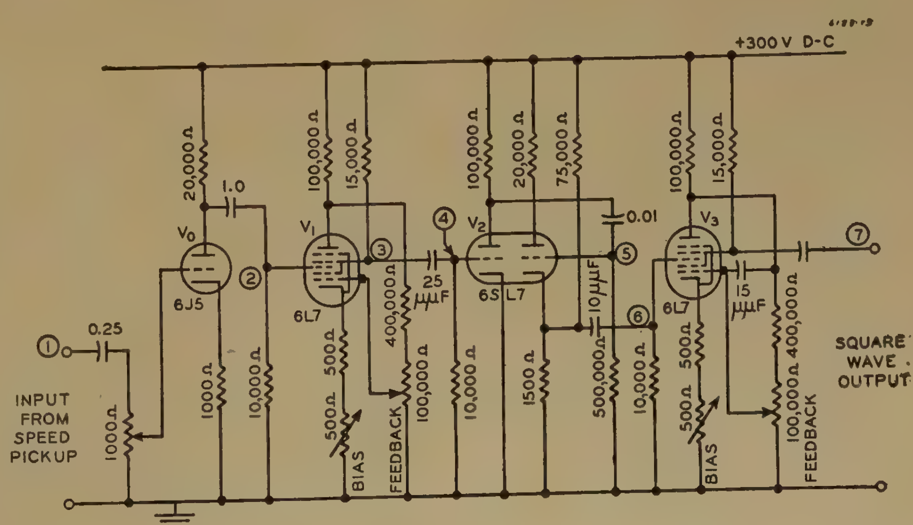
each end of the cylinder housing, a solenoid and adjustable spring assembly is provided to position the armature within the cylinder. When the armature is moved to the right far enough to open the inlet port, high pressure air will flow around the undercut middle section of the armature and through the center outlet to the diaphragm of the steam control valve. The pressure built up in the diaphragm chamber will depend upon how much and how long the inlet port is open. When the armature is returned to the neutral position so that both ports are closed, the output pressure will remain at the built-up value until the armature is moved to the left to open the exhaust port or until the air leaks out around the armature bearing surfaces. Thus by proper positioning of the armature, the output pressure can be set to any value between 0 and 100 pounds per square inch.

Each of the solenoid coils is connected as the plate load of a 6L6 power output tube in a phase-discriminating servo amplifier shown in Figure 8. The two 6L6 plate circuits are energized, 180 degrees out of phase, by the 60-cycle-per-second line via the high voltage transformer T_2 whose secondary is grounded at its center-tap. At zero a-c input to the 6L6 grids, the tubes are balanced to supply equal plate currents to the two solenoid coils. A 60-cycle-per-second error signal applied to the grids such that it is in phase with the plate voltage of the upper tube will be out of phase with the plate voltage of the lower tube; therefore the plate current of the upper tube will increase and that of the lower tube will decrease. This will result in a differential magnetic pull on the valve armature, which will move the armature toward the inlet port to increase the output air pressure. Similarly, if the phase of the input error signal is reversed, the lower tube will conduct more plate current than the upper one, moving the armature toward the exhaust port to decrease the output air pressure.

The current in each solenoid is half-wave rectified 60-cycle-per-second, partly smoothed by means of a 0.5-microfarad shunt capacitor. The remaining a-c component of coil current is utilized to buzz the armature at 60 cycles per second and minimize static friction in the valve bearings. A portion of the second-harmonic cathode voltage is fed back to the input voltage amplifier to provide added stability and an over-all gain control.

The 60-cycle-per-second sine-wave input to the servo amplifier consists of speed error signal (coming in through a "control sensitivity" attenuator and the bat handle switch) in series with air pressure follow-up signal (from the pressure gauge bridge circuit via transformer T_1). The electromagnetic pressure gauge consists of two

Figure 13 The input trigger amplifier



Electrical Anti-icing of Aircraft Windshields

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ASSOCIATE AIEE

PREVENTION of ice formation on the aircraft windshield presents difficult problems. In addition to optical quality, the thermal and structural characteristics of suitable transparent materials, such as glass and certain plastics, must be given careful consideration. Further, an important design criterion for an aircraft windshield is its ability to withstand impact loads such as those which result when a large bird strikes the windshield. Contemporary windshield design practice usually calls for a laminated structure, commonly consisting of a layer of polyvinyl butyral, encased between two plates of tempered glass. The impact resistance of such a construction has been shown to be dependent upon the vinyl temperature, the optimum temperature being of the order of 110 degrees Fahrenheit. Direct correlation thus exists between impact resistance and anti-icing ability in windshield design. The regulations of the Civil Aeronautics Board specify that the windshield and supporting structure of a transport airplane shall be designed to resist the impact of a 4-pound bird when the airplane is flying at sea level cruising speed.[1] It is apparent, therefore, that it may be necessary to supply heat to a laminated windshield continuously, regardless of the prevalence of icing conditions, and further that the amount of heat must be controlled within the limits specified by impact resistance as well as deicing.

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Five schemes or combinations thereof have been advanced for the prevention of ice formation on the windshield:

1. Fluid and wiper.
2. Hot air.
3. Resistance wires embedded in the windshield.
4. Radiant heat (infrared).
5. A transparent electrically-conductive coating.

It is the purpose of this paper to examine the basic considerations for the design of a windshield anti-icing system, and to discuss in some detail the pertinent characteristics peculiar to each scheme. The design will be considered primarily as a windshield ice-prevention problem, with impact resistance as the major secondary consideration.

Heat Requirements

Thermal energy is removed in several ways from a heated windshield which is being flown in an icing environment. Assuming flight through a dense formation of super-cooled water droplets, eff tive anti-icing will be achieved by ma taining the outer surface of the win shield at a temperature above freezi say 35 degrees Fahrenheit. The heat a sorbed from the surface can be group into four basic categories:

1. The loss by forced convection and rad tion.
2. The heat of liquid required to raise t temperature of the water droplets striki the surface.
3. The loss by evaporation of that porti of water required to maintain 100 per ce vapor saturation in the surface bounda layer.
4. The edge losses by conduction to t structure.

A fifth category, kinetic heating,* shou be considered, representing a gain of hea

The relative importance of kinetic hea ing is dependent upon the amount of fr water in the cloud. For a low water de sity and small droplets, the droplets ma be evaporated entirely before they stri the windshield surface. For a high water concentration, water will strike th surface and at once be vaporized so th the surface remains dry; in this case, th kinetic heat equals or exceeds the he losses. Insufficient kinetic heating characterized by a sharp reduction in th temperature of the surface (with a cons quent icing over) unless auxiliary heat supplied. Complete wetting of the su face will be brought about by a relative high concentration of water, and all fiv factors plus auxiliary heat achieve im portance in an anti-icing heat balan analysis.

The relative magnitude of the therm losses for a given airplane and windshie are dependent upon the outside air tem perature, water droplet size, amount water per unit volume, altitude, and ai craft velocity. Two other determinin factors, the surface boundary layer ve locity distribution and amount of wate catch, are influenced by those structur conditions which determine the aero dynamic flow characteristics over th windshield. Briefly consider each pr viously mentioned heat loss.

Loss by forced convection in wet air greater than in dry air because of th higher specific heat of wet air; the radia tion loss is relatively small. The heat liquid required to raise the water tempera ture is dependent upon the amount water catch on the windshield and th temperature differential through which is raised. The loss by heat of evaporatio

* In wet air Prandtl's number approaches unit and kinetic heating becomes proportional to th airplane velocity squared.

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similar coils whose inductances are varied equal amounts but in opposite directions by changes in servo output air pressure. These coils are connected along with the air pressure control dial into a 60-cycle-per-second bridge circuit. By means of the air pressure control dial, the bridge can be balanced to zero at any given operating air pressure. Then any change in pressure will produce a bridge unbalance and a proportional output error voltage, which will energize the servo amplifier to return the air pressure to its original value and rebalance the bridge. If, however, the air pressure control dial is displaced, the servo will stabilize the output pressure at the new value required to rebalance the bridge at the new setting of the control dial.

By means of the resistor R_6 and capacitor C_3 the follow-up voltage at the secondary of T_1 is phased exactly with the speed error signal from frequency meter.

When the bat handle is held down, the speed error signal is disconnected and the servo responds only to the bridge circuit, so that the air pressure control dial can be used for manual speed control.

requires more detailed explanation. As the temperature of the air in the boundary layer over the windshield increases, because of the acquisition of auxiliary heat, it possesses a greater ability to absorb water. If the air at the initial temperature is considered 100 per cent saturated, a lower degree of saturation will exist at the higher temperature. Because of the resulting differential vapor pressure conditions, liquid will be evaporated from the water droplets and the wet windshield surface, and a considerable amount of thermal energy will be extracted in the process. Loss by conduction at the frame is a function of the frame configuration and thermal contact at the glass edge.

The temperature specifications for maximum impact strength have been mentioned. Figure 1, taken from reference 2, indicates a maximum strength when the vinyl plastic is maintained at 110 degrees Fahrenheit, but little deviation from the maximum strength results over a temperature range of from 100 to 120 degrees Fahrenheit. When the vinyl is at a temperature corresponding to that of cabin air, say 70 degrees Fahrenheit, the impact strength based on penetration velocity is reduced by about 45 per cent. This is a serious diminution. The data of Figure 1 refer to a specific windshield and frame design. Other similar designs will have different absolute maximum impact resistance, but the maximum will occur at about the same vinyl temperature.

It is evident that if impact resistance be considered a design criterion of an anti-icing windshield, the thermal requirements will be influenced accordingly. This is particularly true when the time schedule of operation is considered, since adequate impact resistance may require application of heat continuously while the airplane is in flight; anti-icing may be required only at infrequent intervals. The relative heat demand is modified by variations in glass and vinyl thickness, method of heat application, and plane at which heat is developed initially, as well as differences in the flight atmosphere.

Design Considerations

Definite factors influence the adoption and design of a windshield anti-icing scheme:

1. Anti-icing ability.
2. Impact resistance. Proper structural design is involved, possibly supplemented by heat.
3. Weight.
4. Optical properties. Freedom from annoying reflections, adequate light transmission, freedom from optical distortion, and so forth, are necessary.
5. Ease of control. The arrangement must lend itself to a reasonably simple and reliable control system with rapid response characteristics.
6. Transient thermal properties. The system should be able to commence effective anti-icing quickly from an initially unenergized state.
7. Structural factors. Depending upon the source of heat, the structure must be adapted to the primary needs of the system. A simple structural and interior layout is desirable.
8. Fundamental means for ice prevention. Other than fluid deicing, four thermal systems are available. Selection of a system must include consideration of the influence of the deicing system on the functional behavior of other components of the airplane.
9. Reliability and safety.
10. Aerodynamic considerations. The velocity distribution over the windshield surface and the amount of water catch will be determined by the air flow characteristics over the nose and upper front fuselage. Thermal requirements are affected directly.
11. Pilot comfort. A high temperature at the inner windshield surface will result in excessive radiant heat on the pilot. Discharging large amounts of warm air in the cockpit enclosure from a hot air system is likewise undesirable.
12. Maintenance and ease of servicing.
13. Fabrication of windshield. A means of ice removal requiring a panel difficult to manufacture and correspondingly costly is to be avoided. Replacement as well as initial cost would be high.
14. Defogging and defrosting. Collection of fog or frost on the interior windshield surface also must be avoided, the simplest means being to maintain the inner surface above the dew point temperature for the given cockpit air humidity, density, and temperature.

Miscellaneous factors arise from other sources. For example, one air line is considering the addition of a thermal deicing system on a modified Douglas C-54 airplane. Since the line figures a revenue loss of over 1,000 dollars for each day the airplane is out of service, a system which can be installed quickly is highly desirable, even though a weight penalty of several pounds is exacted.

A discussion of the possible methods of supplying heat to the windshield will assist in forming a decision as to the most desirable scheme for a given airplane.

Nonelectrical Methods

Two nonelectrical deicing systems are in common use and will be discussed briefly for comparison with electrical methods. The earliest and most widely used method consists of a mechanical wiper which also distributes an alcohol fluid freezing point depressant over the wiping surface. There are definite limits to the deicing ability, and of course no protection at all is afforded after the supply of fluid is exhausted. The system is, however, simple and reasonably light in weight, and may be applied readily to most structural configurations without modification. Curved

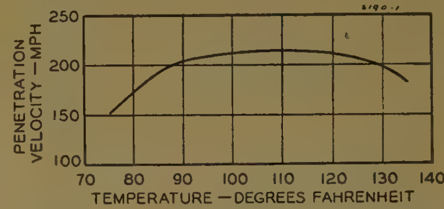

Figure 1. Variation of bird penetration velocity with temperature of a laminated glass and vinyl plastic windshield

windshields present an obvious problem. Other advantages and disadvantages are evident and will not be enumerated.

Hot air anti-icing has received much attention since 1940.[3] The most commonly used arrangement consists of two panes with an air gap of from 0.1 to 0.2 inch. The heated air is introduced at one edge, passes between the two panels, and is discharged directly into the cockpit, over other windows for defogging, or elsewhere. The inner panel is usually of impact resistant construction and the outer designed for maximum heat transfer consistent with adequate strength. One heated air scheme introduces the air to the windshield on the outside surface in the form of an artificial boundary layer.

The hot air system is reasonably light in weight, although the weight of a second panel is appreciable. Optically, the double windshield is poorer than a single pane arrangement because of the multiplicity of light images, both transmitted and reflected, occasioned by the two additional prime surfaces. A distinct disadvantage is the high temperature and mass of air flow required to provide the necessary thermal energy for combatting a severe icing atmosphere. This may overheat any vinyl plastic present, causing permanent small bubbles to form, and also appreciably reduce the impact strength of a laminated panel. Further, the resulting hot inner surface radiates excessive heat to the pilot's face. The glass may crack because of the high thermal gradients. However, the system has been adopted for numerous military and commercial aircraft, largely because of the relative ease in obtaining heated air, either from waste exhaust heat or primary

combustion heaters, compared with the difficulty of procuring sufficient electric power and specially designed windshields.

Embedded Wires

Perhaps the most obvious electrical method of developing heat at the windshield surface is by means of embedded resistance wires. This idea is not new. Rodert describes experiments in which a series of wires were mounted in a space between two glass panels with the intervening region filled with liquid ethylene glycol.[4] The method proved unsatisfactory because of the difficulty in sealing the configuration against leaks of glycol liquid. A windshield of laminar construction with a layer of vinyl plastic sandwiched between two panes of tempered glass conceivably could have wires embedded in either the glass or the vinyl. Because of fabrication difficulties, however, it is impracticable to embed the wires in the glass. One air frame manufacturer has intensively pursued an investigation of panels having the fine wires situated at the interface of the glass and plastic.

The wires must be of a material which can be drawn to a diameter of about one mil or less and still possess adequate tensile strength. The resistance must be correlated with the required deicing power per square foot, wire spacing, and operating voltage. The wire surface should be dull to avoid reflections. Several materials are suitable, such as copper, platinum, copper encased in platinum for strength, beryllium-copper, or nickel. During manufacture the wires are bonded to the vinyl while under tension. Wire buckling and kinking thus is prevented both during panel fabrication and while being heated electrically. The ends are soldered to a bus, usually running horizontally at the top and bottom of the windshield with the wires contained vertically. Visual distraction is less with this arrangement than with the wires mounted horizontally.

A typical design will be described. Beryllium-copper wire one mil in diameter is embedded at or very near the interface between the outer pane and the vinyl. The thicknesses of glass and plastic depend upon the impact strength desired, except that a sufficiently thick front glass pane is required to permit a uniform temperature distribution over the outer face. Three-quarter tempered glass panels 3/16 inch thick, with 1/8 inch to 1/4 inch vinyl are typical dimensions for a windshield 15 inches by 30 inches over-all. Assuming an anti-icing heat de-

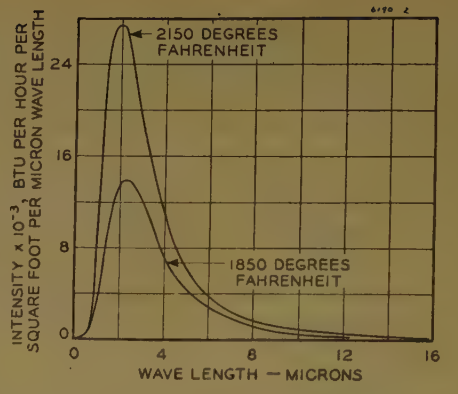

Figure 2. Dependence of radiation intensity upon wave length for a black body

mand of 1,740 Btu per hour per square foot with a five-per-cent loss into the pilot enclosure, 535 watts per square foot must be developed by the heated wires. Mounting the wires vertically, and allowing a one-half-inch-wide horizontal bus at top and bottom, the wires will be 14 inches long. One mil beryllium-copper has a resistance of about 60 ohms per foot, and at 27 volts each wire will dissipate 10.4 watts. Thus 60 wires will be necessary per foot of panel, or a wire spacing of 0.2 inch.

A German windshield, reported by a technical observer of the "Combined Intelligence Objectives Subcommittee" under the British and United States Armies, is described as having 0.015 millimeter (0.6 mil) of copper wire embedded in a plastic middle layer, with a very narrow spacing of the order of 0.05 inch.

The deicing ability of an embedded wire panel should be good, but will be limited by the temperature developed in the vinyl. Exceeding the temperature for maximum impact strength is one consideration; another is a 200-degree-Fahrenheit limit imposed by the tendency of the plasticizer to vaporize above this temperature, forming permanent small bubbles. Values much higher than the critical limit for bubble formation must be developed at the wire surfaces to produce a lower required average vinyl temperature.

The important factors favoring hot wire windshields are the light weight associated with operation on 28 volts, their reliability, and the efficient manner in which the developed heat energy is dissipated.

The disadvantages of the scheme are fundamental and serious. Optically, there are two deterrents. First is the necessity for looking through the closely spaced fine wires with the concomitant eye fatigue. Viewed from a distance of two feet, with eyes focused on objects beyond the windshield, the visual distraction is slight but not insignificant. Second, and of greater importance, is the optical distortion caused by the local heating of the vinyl sheath about each wire. The resulting localized variation of the index of refraction will distort severely objects viewed through the panel. The extent of distortion is proportional to the temperature gradient, and provision of sufficient heat for anti-icing under moderate to extreme conditions may render the windshield optically useless. This condition is tempered by virtue of an outside visibility of zero during icing conditions anyway, but the significant time required for the distortion to fade away after the panel is de-energized may be serious. The distortion produced by the heat necessary in clear air for interior defogging and for impact strength further establishes the seriousness of the situation.

The necessary construction renders the windshield difficult and expensive to manufacture, and the deicing ability is limited by local overheating of the vinyl. If the necessity for supplying heat in clear air conditions (for impact strength and defogging) or the inadvisability of so doing (because of distortion) are disregarded, a complication is introduced by the thermal lag of the system. The time which elapses between the application of power and the inception of ice removal could be serious in an emergency.

Infrared Radiation

Radiant thermal energy directed on a solid body will be absorbed to a varying degree. Most solids are sensibly opaque to infrared radiation, but because of the characteristic similarity between energy in the visible spectrum and in the near infrared spectrum, visually transparent substances such as windshield constituents will pass a certain portion of the energy entering the surface. Because of surface reflection, only a part of the radiation impinging on the surface will enter any material.

The amount of energy reflected or absorbed by a receiver is a function of the wave length of the radiations emitted by the source.[5] The visible spectrum is included within the range of from 0.35 to 0.8 micron. The infrared spectrum is divided arbitrarily into the near infrared spectrum ranging from 0.8 micron to around 25 microns, and the far infrared spectrum ranging above 25 microns. In general, thermal radiation at all wave lengths is emitted by hot solid bodies, the total emission depending upon the temperature, the material, and the condition of the surface. Typical radiation

patterns for a body at two temperatures is shown in Figure 2. A small fraction of the total energy is in the visible range but the greatest portion is in the near infrared spectrum. At higher temperatures the peak of the curve is shifted to a lower wave length, the position of the maximum being inversely proportional to the absolute temperature.

The dependence upon wave length of the total energy absorbed by an optically transparent receiver is indicated in Figure 3. Ordinary glass and vinyl plastic readily pass the visible frequencies. The very near infrared spectrum is largely passed by the glass, but strongly absorbed by the vinyl. Thermal radiation at frequencies above 2.8 microns is absorbed by both materials. Special glasses are available which are opaque to visible radiation, but which transmit most of the infrared rays from one to four microns. Conversely, others transmit most of the visible rays, but strongly absorb all wave lengths above 0.8 micron.

An anti-icing scheme using infrared rays is apparent. A grid consisting of electrically heated wires or a series of specially constructed radiant heat lamps are practical as a source. A polished metal reflector will assist in concentrating or distributing the energy on the windshield. An infrared filter may be incorporated, depending upon the amount of visible light developed (a direct function of source temperature) and the necessity for eliminating it. Directing the radiation on the windshield will cause it to rise in temperature, and if sufficient energy is supplied, effective ice prevention may be accomplished.

Most of the heat is developed within the windshield because of the relatively strong absorptive characteristic of the vinyl. The proportional amount for a specific case may be determined from Figure 3. The heat then is transmitted by conduction to the outside. The amount of energy passing on through the assembly may be reduced by using a heat absorbing glass as the outer pane, thus collecting more heat energy in this plane. However, radiation energy passing completely through the windshield will be absorbed by ice or water which may have collected on the surface. Laboratory tests show that a layer of ice so irradiated tends to "mush-up" and melt throughout the layer, a phenomenon readily explained by the penetrating power of the radiation.

A similar layer melted only by heat conducted through the glass displays a tendency to melt away at the glass surface, leaving an ice bridge over the windshield but separated from it by a small air or water filled space.

Suitably locating the source of the infrared rays is one of the major problems presented by this system. Pilot vision must not be impaired, but a reasonably uniform distribution of heat must be secured. These conditions are difficult to meet except for highly sloping windshields, since the incident radiation should strike the windshield at as near 90 degrees as possible. Any solution for windshield configurations similar to the DC-3 or DC-6 is apt to be awkward, and require considerable skill in overcoming the inherent spatial problem. The ratio of source space required to power emanated is of the order of one to two cubic inches per watt. The applicability of radiant heating thus is seen to be greatly dependent upon the structural configuration of the cockpit.

One answer to the space question is to locate the source outside of the airplane. Some form of retractable mounting for hermetically sealed source could be employed in conjunction with a filter to eliminate the visible light. The slope of the windshield may make it desirable to mount the source above the windshield rather than on the nose. This or other factors combined with the high space-to-power ratio again emphasize the fundamental awkwardness of the scheme based on infrared radiation.

Advantages of a radiant heating scheme are light weight and ability to operate on 28 volts. A special windshield design is unnecessary, and the appropriate plastic temperature for satisfactory impact strength may be maintained, although this temperature will not be synonymous with the all anti-icing heat requirements. Indeed, because of the absorbtive capacity of the vinyl in the near infrared spectrum and its low thermal conductivity, an inside source may overheat seriously the inner plane of plastic in an attempt to transmit sufficient heat to the outer glass.

Major disadvantages are the cumbrous installation demands, the difficulty of obtaining even heat distribution over the windshield, and the fire hazard inherent in operating the source at a suitably high temperature. The over-all energy efficiency is poor because of heat losses at the source by convection and conduction plus the loss in a filter should one be necessary. There are also practical limits to the amount of heat energy that can be developed by the source, and to the amount the windshield can absorb without undue temperature gradients within the glass and vinyl sections. The former condition is more of a limitation than the latter, and is related to the space and safety requirements and thermal efficiency. Either factor may make it impossible for the infrared radiation system to provide sufficient protection against the expected icing environment.

Transparent Electric Conductor

A transparent electrically conductive coating on glass, a coating of unique properties, recently has been developed and introduced by a major glass manufacturer.* The importance and application of such a coating for windshield deicing is at once obvious. A coating of this general description is not new; a film of high electrical resistance has been used in the past on glass surfaces for the prevention of electric charge collection, or electrostatic shielding. The new coating, however, exhibits the very important advantages of relatively low resistance and indestructibility. Instead of resistivities of the order of megohms, the new so-called *NESA* coating, can be produced with a resistivity as low as 100 to 120 ohms per square, this being the resistance measured between opposite sides of a square sheet of any linear dimension. Some idea of the thinness of the *NESA* coating is gained by considering that a film of copper having a resistivity of 100 ohms per square would be only seven billionths of an inch thick (assuming that the dimensional resistance properties are maintained for very thin sections, which may not be the case). This is about 0.03 of one per cent of the wave length of light.

The exact nature of the *NESA* film is not known to the author, but its physical properties and limitations can be described. The basic material is applied to a glass surface and processed at relatively high temperatures. It cannot be applied to such materials as Plexiglas (because of the high processing temperature, and possibly other factors). The film has excellent abrasive resistance properties, being of the nature of the glass itself. The coated glass may be partially tempered, but a high temper presents some difficulties. A laminated panel can be made with the coating next to the vinyl plastic if so desired. A *NESA* coating within a laminated panel will introduce a loss in light transmission of around four per cent largely by reflection loss caused by the difference in the index of refraction of the glass or vinyl and the *NESA*. A coating on the outer glass surface will double this loss and occasion more objectionable reflections. The index of refraction of the

* The Pittsburgh Plate Glass Company.

NESA is approximately two. When viewed by transmitted light, the panel imparts no noticeable reflections or color. By reflected light, the reflections from the NESA surface are sharp and possess a greenish or reddish caste. The coating appears mottled under close inspection and formed of small splotches reflecting red and green.

Electrical connection is established by means of a narrow opaque bus applied as a coating to the surface of the NESA film and usually running lengthwise along opposite edges of a panel. A copper strip may be soldered to the bus and brought out to serve as a terminal connection point.

The uniformity of the NESA coating on an aircraft windshield is important, since a nonuniform film will produce localized hot and cold spots. Several large laminated panels, with the coating either on the outer glass surface or next to the vinyl, have been laboratory tested for thermal variance. The results show a maximum deviation from the average of about plus or minus three degrees Fahrenheit for spot checks at points two inches apart over the entire windshield surface. These data were obtained with the horizontally mounted panel in still air with a steady state average surface temperature of approximately 100 degrees Fahrenheit.

A windshield has been designed for a pressurized Douglas airplane which is expected to provide freedom from icing under the most severe anticipated conditions at any operating speed or altitude of the plane. Further, impact resistance of significant excess over the minimum requirement may be incorporated at all times. The panel is approximately 13 inches wide, 28 inches long at the top and 31 inches long at the bottom. A maximum power of 3,100 Btu per hour per square foot is to be dissipated by the NESA film which has a resistivity of 120 ohms per square. The heat requirement was computed by methods previously described as that required to anti-ice for the severest expected atmospheric conditions at minus ten degrees Fahrenheit, sea level, and 270 miles per hour, and assuming that the outer surface of the windshield is maintained at 35 degrees Fahrenheit. The average vinyl temperature will be controlled at 110 degrees plus or minus ten degrees Fahrenheit for greatest impact strength, and will not exceed 125 degrees Fahrenheit under this maximum power demand condition.

First considerations in the design were the location of the NESA layer, and the glass and vinyl thicknesses. For the most

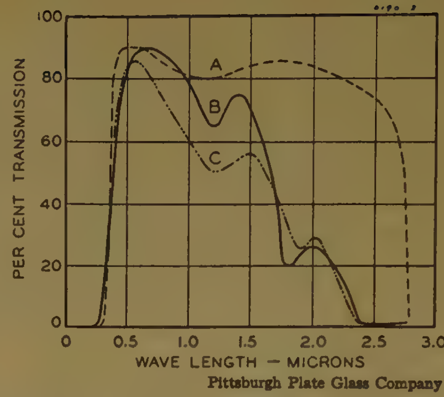

Figure 3. Transmission characteristics as a function of wave length

A—Polished plate glass, 6.17 millimeters
B—Vinyl plastic, 3.1 millimeters
C—Laminated panel, 0.120-inch vinyl plastic encased between two 7/64-inch glass panes

rapid response and most effective deicing, the coating should be put on the windshield outer surface. This, however, introduced the problem of protecting the bus from abrasion, a safety hazard because of the exposed coating, and necessitated other means, such as another interior coating, for maintaining a warm vinyl and for defogging. The best compromise was a NESA coating on the inner surface of the outer pane. The power of 3,100 Btu per hour per square foot will be conducted through the outer pane at a negative temperature gradient of 465 degrees per inch, neglecting edge conditions. A glass panel 3/16 inch thick will transmit this thermal energy at a temperature difference of 87 degrees Fahrenheit, thus requiring 122 degrees Fahrenheit at the vinyl-glass interface, a figure which is approximately the upper limit for maximum impact strength. A thinner front pane would be satisfactory thermally and more desirable from the standpoint of rapid transient response, but strength and glass temper considerations lead to the selection of the greater thickness. The same thickness was adopted for the inner pane, and 1/8 to 3/16 inch adopted for the vinyl. The temperature difference across the vinyl under operating conditions will be only a few degrees.

The described windshield incorporates a 5/16-inch bus running horizontally at the top and bottom edges behind the mounting frame. The heated area is 2.65 square feet requiring 2.4 kw at 3,100 Btu per hour per square foot. This amount of power will be developed by 350 volts for a coating resistivity of 120 ohms per square.

Control may be effected by maintaining a vinyl temperature appropriate for high impact resistance. At 110 degrees Fahrenheit nearly all icing environment can be countered; for the most severe conditic the control setting can be increased 120 or 125 degrees Fahrenheit. Only minor sacrifice in impact strength will introduced by maintaining the tempe ture in the upper range, and full anti-ici ability will be provided at all times. if sufficient strength is developed at lower temperature, say 70 degrees Fahre heit, control could be maintained at t point with a corresponding improveme in pilot comfort in a warm cockpit. Se eral schemes for control have been a vanced which will not be enumerat here. The design will depend large upon the following three factors:

1. The source of electric power and type generating equipment.

2. The type of temperature sensing eleme incorporated with the windshield.

3. The transient thermal response of t windshield and associated components.

At the maximum flying speed in co clear air, sufficient power must be su plied to the windshield to maintain t optimum vinyl temperature. The pr viously quoted figure of 3,100 Btu p hour per square foot will suffice to mai tain a vinyl temperature of 120 degre Fahrenheit for an outside air temperatu of minus 70 degrees Fahrenheit at 27 miles per hour.

Provision for stand-by in case of co trol or generating equipment failure ma be obtained by using two control units an two alternators or inverters. Alternatin current is the most logical type of powe because of the high voltage requiremen and absence of disturbing unidirection fields. One generator may be of a capac ity that is sufficient to do the whole jo with a second machine for stand-by; o two generators of half capacity may b used with a switching arrangement t allow application of power to the pilot' windshield system only, should his gen erator or control fail. Thus at least ha of the pilot's enclosure could be kept ice free in an emergency, and for light icin conditions the one generator could deic the entire windshield system.

The transparent electrically conductin coating of low resistance appears to be a ideal solution to the windshield icing an impact proofing problem. The maximu amount of heat required for the wors conditions may be supplied easily an used efficiently. The system lends itse readily to control both from anti-icing an impact resistance considerations. NESA coated windshield possesses satis factory optical characteristics at all times It is relatively simple to manufacture, no subject to failure in operation, and allege

Moisture Equilibrium Between Gas Space and Fibrous Materials in Enclosed Electric Equipment

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THE USE of an inert gas in enclosed electric equipment has increased considerably in recent years. In transformers an inert gas is used in the space provided above the transformer fluid to allow for its volumetric expansion with temperature rise. There the function of the gas is to provide an inert medium by which the pressure at the surface of the transformer fluid is kept slightly above atmospheric. Thus the fluid is not exposed to the moisture and oxygen of the air, as it is with the so-called "open-breather" or "oil-conservator" types unless the latter communicate with the outside atmosphere through chemical agents. In gas pressure cables, however, the inert gas is used at pressures up to approximately 15 atmospheres in order that the cables may operate successfully at electric stresses that are higher than otherwise would be possible.[1-3]

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The purpose of this paper is to call attention to a phenomenon that occurs in enclosed electric equipment in which both an inert gas and fibrous material are present. The systems involved tend to establish an equilibrium between the moisture content of the fibrous materials and that in the gas space. Consideration, therefore, should be given to the moisture content of the fibrous materials present as well as that of the gas itself. A quantitative method is suggested for predetermining the equilibrium condition for both unimpregnated and impregnated fibers. It is believed that use of the method may assist in the solution of manufacturing and operating problems.

Field Experience and Data

The increasing of the moisture content of dry nitrogen by the transfer of moisture from fibrous materials has been observed in The Detroit Edison Company in transformers and in the 120,000-volt gas-pressure cable.[1-3] This increasing of the moisture content of the gas can be of practical significance only if the dew point of the gas is reached at local cool spots over considerable periods of time, thus causing progressive transfer of moisture from the warmer fibrous materials, and if the resulting condensate drops upon a vulnerable part of the equipment. Any condensation that takes place as a result of an over-all cooling of the equipment or of a cool spot of short duration is of no significance because the water content of the gas is relatively low.

During severely cold weather, it is probable that some condensation occurs in all of the transformers that have been investigated. Moreover, it is probable that, in certain transformers, condensation occurs during the cool part of the day nearly every day of the year. Apparently the condensate usually drops in locations where it does no harm. Condensate has been found, however, on the terminal board inside transformers. Whether or not service failures actually have been caused by condensate on vulnerable parts is unknown, but there seems little question that such condensation constitutes a potential source of trouble.

In the case of the 120,000-volt gas-pressure cable, in which the paper-insulated unsheathed cables rest on the bottom of a steel containing-pipe, nearly all locations are vulnerable to attack by liquid water should condensation occur at local cool spots. Such cool spots might occur at a terminal, a length that crosses

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to be free from deterioration caused by aging, sunlight, overheating, and abrasion. During operation the inner surface of the windshield is not excessively hot, which improves pilot comfort by minimizing the radiant heat. The installation should prove reliable and safe. Structurally, a clean design should result, the only complication being the means for attaching the power wires and control elements.

There are disadvantages to the scheme, however. Most important is the weight, which may be high because of the necessity for using high voltage alternating current delivered by inverters or engine driven alternators. The control system adds significant weight, although a unit to effect the same order of regulation in other ice prevention schemes would be of comparable weight. Inability to operate the *NESA* film directly from the ship's 28-volt power supply and provide the necessary power is a disadvantage. The requisite high voltage introduces a safety hazard, although placing the *NESA* coating next to the vinyl will tend to mitigate the danger. Windows or shields made of plastics such as Plexiglas cannot be deiced or defogged by this method since the *NESA* coating cannot be applied to their surface, at least not by present techniques.

Conclusions

1. Aside from anti-icing, an additional prime reason for heating aircraft windshields is evident. For a given weight and construction, vastly improved impact resistance can be obtained by using a laminated panel with the vinyl plastic maintained at about 110 degrees Fahrenheit.

2. The method of heating a windshield will depend greatly upon the structural configuration of the cockpit and of the windshield system. Availability of heated air and (or) electric power are equally important factors.

3. Use of the transparent electrically conductive coating on glass is the most desirable electrical method for providing heat at the windshield. Indeed, unless operating experience proves otherwise, this method appears to be suited ideally for the purpose of windshield anti-icing. For most airplanes, it ultimately should supplant other presently known systems.

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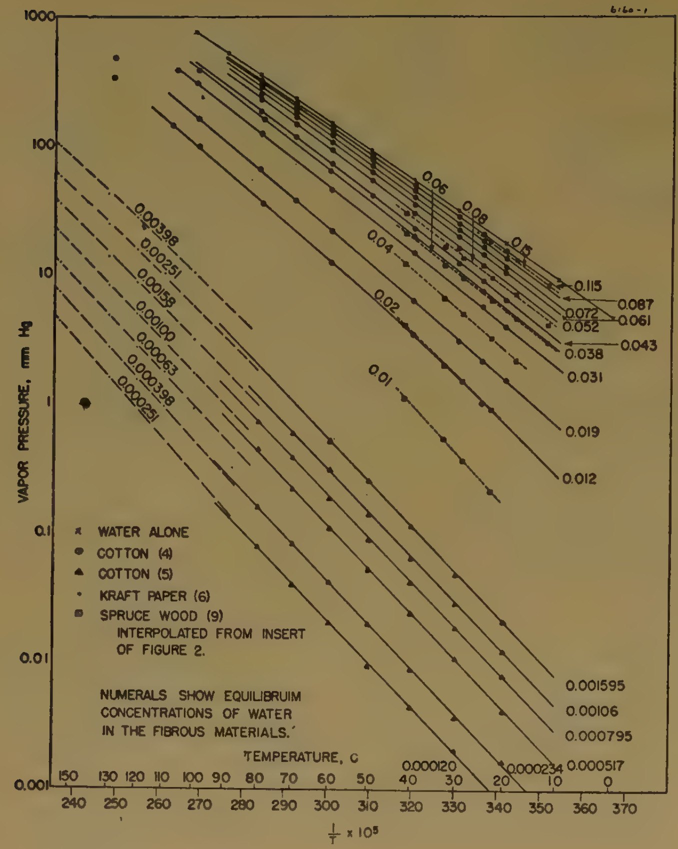


Figure 1. Vapor pressure of water alone and in fibrous materials; data from literature

a canal, or a location in proximity to a water main. For this reason when it was found during the progress of the installation that the moisture content of the dry nitrogen gas, which was introduced immediately after the cable-pulling operation, was beginning to increase, steps were taken co-operatively by the manufacturer and user to determine the cause and corrective measures necessary. These efforts were completely successful as demonstrated both by the data herein contained and by the performance of this circuit since it was energized in December 1941. The data concerning this cable are presented in this paper only because they indicate that a predictable equilibrium exists between the moisture content of the fibrous material, even when impregnated, and that of the gas space.

As has been described elsewhere[1,2] the 120,000-volt gas-pressure cable is of preimpregnated-tape construction; that is, the cable was taped in an air-conditioned room with tapes that previously had been impregnated and from which the excess impregnant had been removed. Over the taped insulation two layers of metallized-paper shielding tapes were applied and over these a thin bronze tape intercalated with an impregnated muslin tape. This was followed by two layers of paraffin-impregnated canvas* to provide mechanical protection during installation. After a spiraled skid wire was wrapped over the canvas and the assembly coated with petrolatum for lubricant during installation, the cable was enclosed in a temporary lead sheath that was stripped within a few feet of the entrance tube to the pipe line as the cable was being installed.

When the dry nitrogen, which was introduced into the steel pipe line with the cable and held at slightly over atmospheric pressure by temporary plugs, began to show the presence of moisture, the source of the moisture was sought. The pipe was known to be dry. No water was found in the petrolatum grease but from 2.4 to 5.0 per cent of water** was found in the impregnated muslin and canvas tapes,* based upon the weight of dry fabric. The average for 14 samples was 3.3 per cent. Two steps were taken to minimize the moisture content of the muslin and canvas tapes.

1. The manufacturer modified his process in such a manner that the moisture content of the muslin and canvas tapes on further shipments of cable was reduced to an average of 1.67 per cent as judged by tests on six samples in which the content varie[d] from 1.0 to 2.0 per cent.

2. Evacuation of the completed line, whic[h] is a normal procedure in the installation o[f] this type of cable, was extended for eigh[t] days while the cable was heated by con[ductor] heating to give a tape temperatur[e] of between 40 and 42 degrees centigrade. The water removed was collected in a car[bon]-dioxide-alcohol trap and measured. From the average water concentration and the weight of water removed, the aver[age] moisture content of the protective tape[s] as the line was placed in operation was cal[culated] to be 1.73 per cent. Presumably most of the water that was removed cam[e] from the wetter tapes.

In order to determine whether thes[e] efforts to prevent the possibility of con[densation] were successful, periodic deter[minations] were made of the moistur[e] concentrations† or the dew points of th[e] gas space and of the temperature a[t] various points along the line. Durin[g] the first year of operation the dew point[s] rose until late summer when at the mor[e] moist locations, which contained cabl[e] not given the special factory treatment they were above the winter operatin[g]

* This type of construction no longer is used by the manufacturer and hence it does not follow that dew point problems are inherent in other gas-pressure-cable installations.

** As determined by Dean-Stark apparatus and boiling toluene.[4]

† Originally the moisture concentrations were de[termined] gravimetrically at atmospheric pressur[e] and the results calculated to line pressure. Late[r] dew points were determined at line pressure using [a] dew-point apparatus.

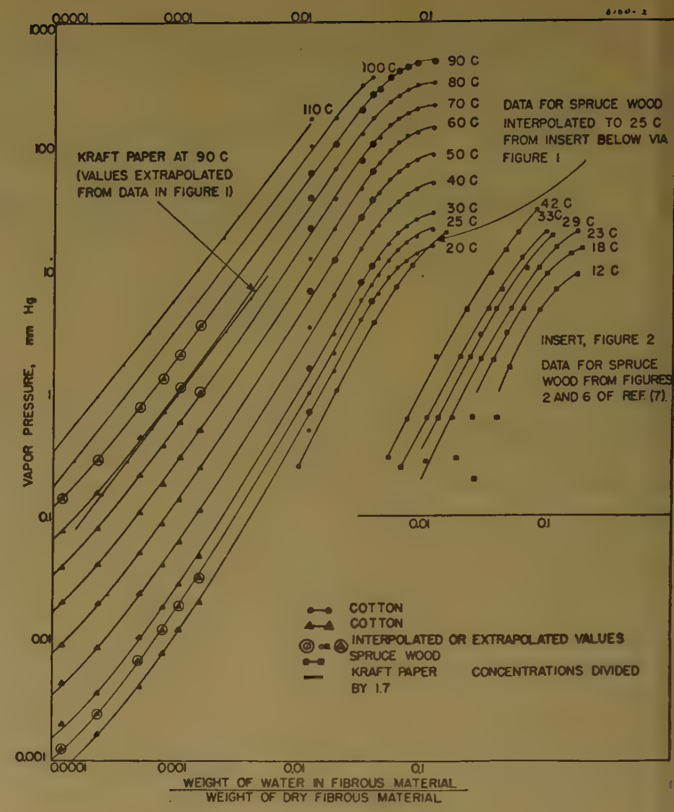


Figure 2. Equilibrium between water con[centrations] in fibrous materials and vapor pres[sure] at selected temperatures

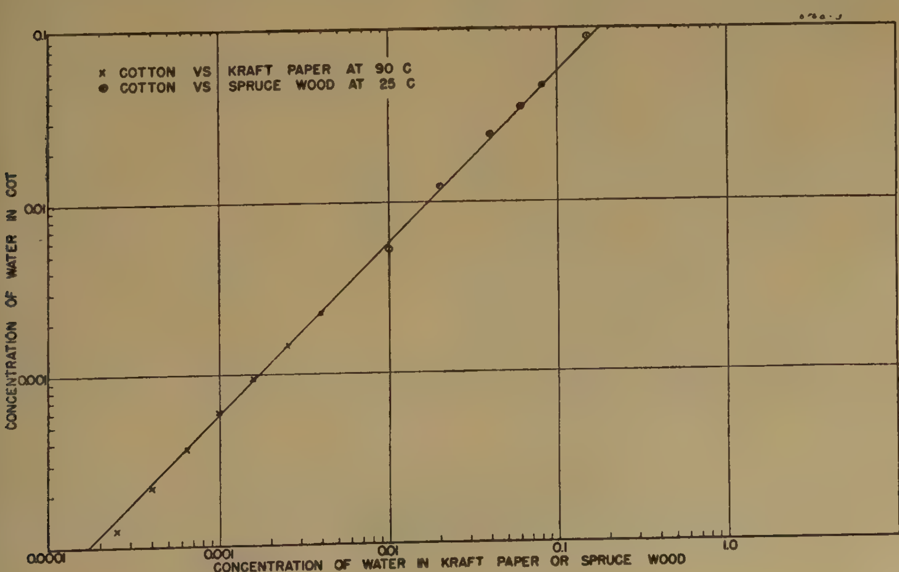
temperature of the pipe line. In winter, however, the dew points fell to values safely below the winter operating temperature.

Inasmuch as the rise and fall, with temperature change, of the vapor pressures that corresponded to the several dew points or moisture concentrations followed the vapor pressure rule for liquids, within experimental error, it became evident that an equilibrium existed between the moisture content of the gas space and that of the tapes. This behavior, coupled with the change in dew points of transformers with temperature change, led to a search of the literature to determine whether quantitative data existed concerning this equilibrium for the fibrous materials generally used in the electrical industry. Literature on the subject revealed that although there are many articles describing the water concentrations of various fibrous materials under various relative humidities, only a few show the effect of temperature change. Fortunately two such sets of data were available for cotton and one each for kraft paper and spruce wood. From these data an equilibrium chart has been prepared that is useful for predicting and interpreting the behavior of the moisture within electric equipment containing fibrous materials and a gas space.

Preparation of the Equilibrium Chart

Figure 1 shows the data as obtained from the literature. In this figure the vapor pressures, on a logarithmic scale, are plotted against the reciprocals of the absolute temperatures on a linear scale. The top curve is for water alone. The long solid curves below the top curve represent the data of Urquhart and Williams[5] for cotton containing over one per cent of water. The data for each concentration are represented as a continuous curve instead of as the two intersecting straight lines used by the original authors. As thus represented, the curves for the higher concentrations are nearly parallel to, and have nearly the same curvature as, the curve for water. With diminishing concentration of water, the curves become more divergent and their slopes approach that of the curve for ice. This curve for ice is not shown in Figure 1. The solid curves at the lower right hand corner of the figure represent the data of Neale and Stringfellow[6] for cotton of low moisture content. These curves are straight, parallel lines having the slopes of the vapor pressure curve for ice. The curves that appear to extend as broken lines from the curves representing Neale and Stringfellow's data represent the data of Houtz and McLean[7] for kraft paper of low moisture content at elevated temperatures. The data for the samples having the higher moisture concentrations are best represented by curved lines whereas the data for the samples of lower concentrations fall along straight lines as shown. The latter are not parallel with each other. For the data of Urquhart and Williams, the slopes of the curves become steeper as the moisture concentrations of the samples represented diminish. The fourth set of data is that of Pidgeon and Maass[8] for spruce wood. These data appear in the same region as those of Urquhart and Williams. Because the temperature range for the investigation on spruce wood was narrow, it is not clear whether the data are best represented by straight or curved lines. The slopes of the lines are approximately the same, however, as those of the adjacent curves for cotton.

Figure 3. Comparison between the concentration of water in cotton and that in kraft paper or spruce wood required to give the same vapor pressure at the same temperature



In using the four sets of data to prepare a general equilibrium chart, no effort was made to distinguish between absorption and desorption data. The difference, where shown by the original authors, is small as compared with the effect of temperature change, which is the primary concern of this paper, and probably is also small as compared with the effect of the previous history of the samples.[9]

The data for the equilibrium chart for cotton were obtained essentially by interpolation, for selected temperatures, of the vapor pressures that correspond to chosen concentrations of moisture. The principal object of the interpolation, of course, was to bridge the gap between the data of Urquhart and Williams and those of Neale and Stringfellow in the concentration range between 0.0016 and 0.012 gram of water per gram of dry cotton. Interpolation was carried out by means of Figure 2, where the vapor pressures and concentrations of water in cotton are both plotted on logarithmic scales for selected temperatures to give the long S-shaped curves that represent both the data of Urquhart and Williams and those of Neale and Stringfellow. In order to obtain the 90-degrees-centigrade curve of Figure 2, the data of Neale and Stringfellow were extrapolated from 80 to 90 degrees centigrade by extending the straight lines as shown in Figure 1.

Because only one set of data was available for kraft paper‡ and that at high temperatures only, interpolation similar to that described for cotton could not be carried out. Instead, the curves in Figure 1 for the data of Houtz and McLean were extrapolated downward from 100 to 90 degrees centigrade, and the resulting values of vapor pressure were plotted against their respective concentrations in Figure 2. The straight line that resulted when the extrapolated values for kraft paper at 90 degrees centigrade were plotted on the double-log scale of Figure 2 lies slightly below the curve for cotton at 80 degrees centigrade as shown. Next, it was reasoned that inasmuch as kraft paper often is made from spruce pulp, possibly the equilibrium in kraft paper and spruce pulp might be related. For this reason the data of Pidgeon and Maass,[8] shown in the insert of Figure 2, were in-

‡ Vincent and Simons[10] evidently investigated the field over a wide range of temperatures and moisture concentrations but unfortunately did not include in their publication data suitable for preparing an equilibrium chart.

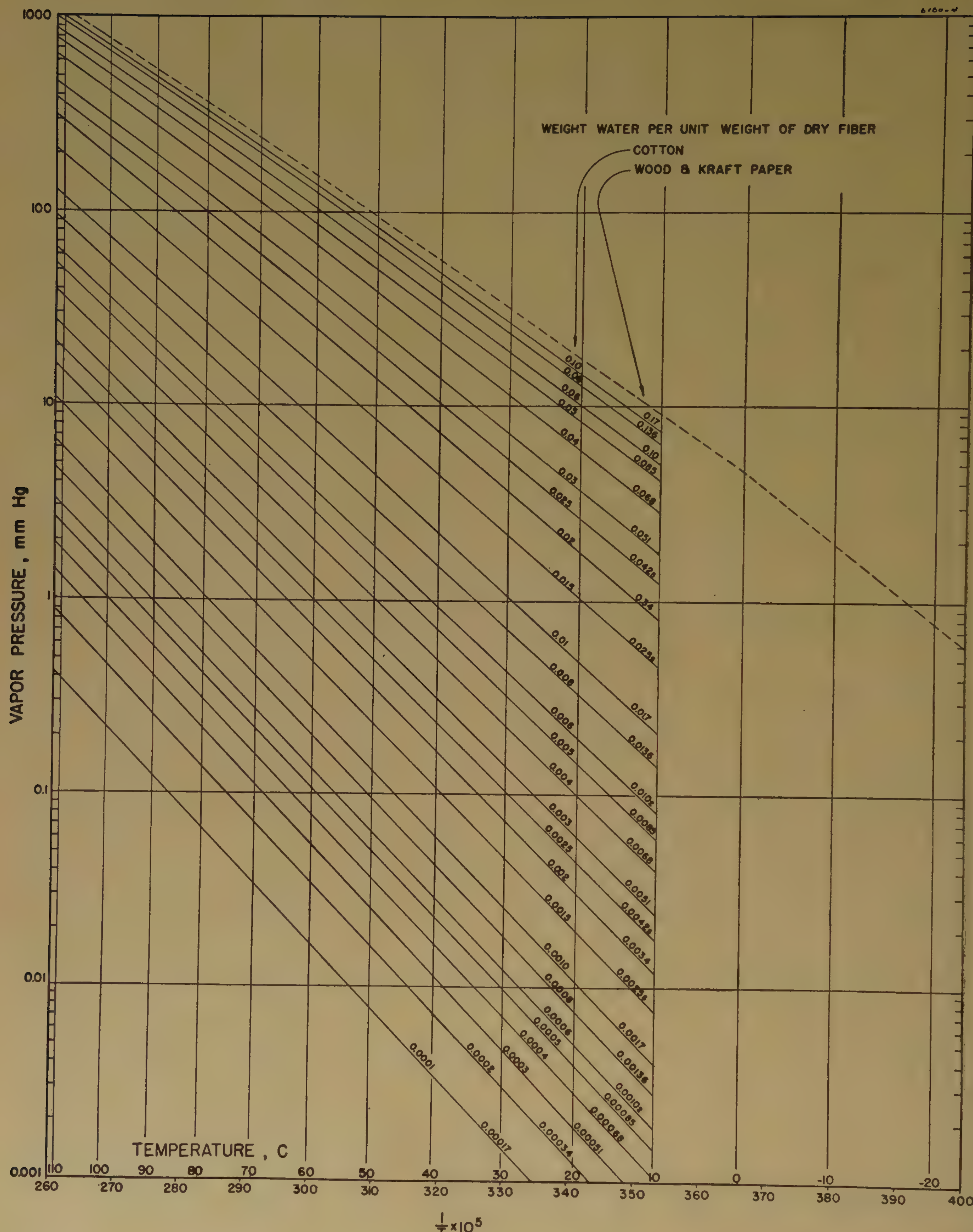


Figure 4. Equilibrium chart showing the relationship between the water concentration in fibrous materials and the vapor pressure of water in space at temperatures between 10 and 110 degrees centigrade

terpolated by the use of Figure 1 to 25 degrees centigrade. Finally, the concentrations of water in cotton required to give the same vapor pressure at 90 degrees centigrade as are given by selected concentrations of water in kraft paper and those concentrations required to give the same vapor pressure at 25 degrees centigrade as are given by selected concentrations of water in spruce wood were plotted on a double-log scale as shown in Figure 3. The straight line through these points indicates that the relationship between the water concentrations that are required to produce the same vapor pressures in cotton and in either kraft paper or spruce wood is simply the ratio 1 to 1.7.

By dividing the concentrations of water in kraft paper, as given in Houtz and McLean's data, by 1.7, an estimate was made of the vapor pressures that would result from low concentrations of water in cotton at 100 and 110 degrees centigrade. These data were combined with the experimental data of Urquhart and Williams in that temperature range to produce the top two curves of Figure 2.

The equilibrium chart was prepared by plotting the vapor pressures for cotton, as given from the curves of Figure 2 for selected concentrations, on a logarithmic scale against the reciprocals of the absolute temperatures, similarly to Figure 1. This chart is shown in Figure 4. All of the curves are extrapolated from 20 to 10 degrees centigrade. Those for concentrations above 0.03 gram of water per gram of dry cotton also are extrapolated for temperatures above 80 degrees centigrade. For this extrapolation, the gentle curvature of the vapor pressure curve for water, shown as the broken line constituting the top curve, was followed.

The equilibrium concentrations of water in cotton that were selected for the preparation of the equilibrium chart are shown by the numerals of each curve in the left-hand column. The numerals in the right-hand column represent the equilibrium concentrations of water in kraft paper or spruce wood, as obtained by multiplying the respective values for cotton by 1.7.

Effect of an Impregnant

Although the practical value of an equilibrium relationship, such as that described in this paper, has been recognized previously for unimpregnated fibrous materials,[10] there seems to be no record of its application to impregnated fibers. Indeed, to many who have tried to dry impregnated fibers that have absorbed moisture, it might appear that impregnated fibers behave entirely differently from unimpregnated ones. The data for the paraffin-impregnated cotton protective tapes of the 120,000-volt cable previously described provide a comparison between the behavior of commercially made American impregnated-cotton tapes and that of the two sets of laboratory-prepared unimpregnated-cotton fibers made in England.

The comparison between the impregnated and the unimpregnated cotton samples is given in Figure 5. The solid lines in this figure, and the numerals arranged in a column upon them, are reproduced for cotton from Figure 4. The crosses represent the average conditions at three locations at which the higher moisture concentrations were found in the gas space as the average temperature of the pipe line rose and fell with seasonal and load change. The low-temperature point was determined while the line was not in operation. The circles represent the average conditions at two locations at which the lower moisture concentrations were found. It is evident that, regardless of whether the fibers are impregnated or not, the change of the vapor pressure of water in the gas space is essentially the same function of temperature change. It is also evident that although equilibrium had not been reached when some of the data on the cable line were obtained, especially those taken in September 1942, the moisture concentrations in the cotton, as indicated by the chart for different dates and temperatures, are nearly the same. Even including the September 1942 data, the indicated concentrations ranged only between 2.2 and 2.6 per cent for the tapes in the more moist regions and between 1.2 and 1.5 per cent for those in the drier locations. Data taken during the initial operation of the cable in the winter of 1942, however, fall far below the equilibrium values shown.

It does not necessarily follow, that because the vapor pressure changes in the same manner with the temperature change regardless of whether the cotton is impregnated or unimpregnated, that the equilibrium concentration of water in the cotton is the same in impregnated cotton as in unimpregnated. Naturally it was impractical to obtain samples of the tapes in the completed 120,000-volt cable line after the evacuation treatment in order to determine whether the impregnant alters the equilibrium. Nevertheless, a means for obtaining a qualitative answer exists. As previously stated, the average concentration of water in the impregnated cotton tapes of the completed line, as calculated from the analytical results and from the quantity of water removed, was 1.73 per cent. The points indicated by dots in Figure 5 show the concentrations of water in cotton at each of the 30 joints at which the samples were taken. These were determined with the equilibrium chart from the average temperature of the line and the vapor pressures that correspond to the dew points that were determined in September 1944. The values, which because they were determined by the chart are for unimpregnated cotton, ranged between 0.9 and 2.65 per cent and averaged 1.68 per cent. The agreement between 1.73 and 1.68 per cent is much closer than the analytical methods warrant. The agreement indicates that it is safe to use the equilibrium chart for impregnated cotton at least in the solution of practical problems in which an error of perhaps plus or minus two in the second significant figure can be tolerated.

No data are available to show whether impregnation with insulating oil changes the equilibrium temperature for kraft paper. The data for the paraffin-impregnated cotton support the belief that although the impregnant greatly slows down the diffusion process by which equilibrium is attained, the equilibrium concentration of water in a fibrous material is not affected greatly by an organic impregnant. It seems probable that the equilibrium conditions shown for unimpregnated kraft paper in Figure 4 will apply approximately for oil-impregnated kraft paper as well.

Use of the Equilibrium Chart

The following examples are intended to illustrate a few of the uses that may be made of the equilibrium chart in the electrical industry. The first three of these examples concern cable of the type discussed.

Example 1

Did the precautionary treatments of protective tapes (previously described) effect any substantial reduction in the dew point of the gas?

This question is answered by the information in Figure 6, which also shows the method of using the equilibrium chart. As previously stated, before the drying treatments the maximum moisture concentration was 5.0 per cent, and the average 3.3 per cent. After the factory treatment on the part of the cables and evacuation of the entire line, the maximum concentration as determined from the dew points was 2.65 per cent and the average concentration from both the dew point measurements and the calculated residual water concentration was

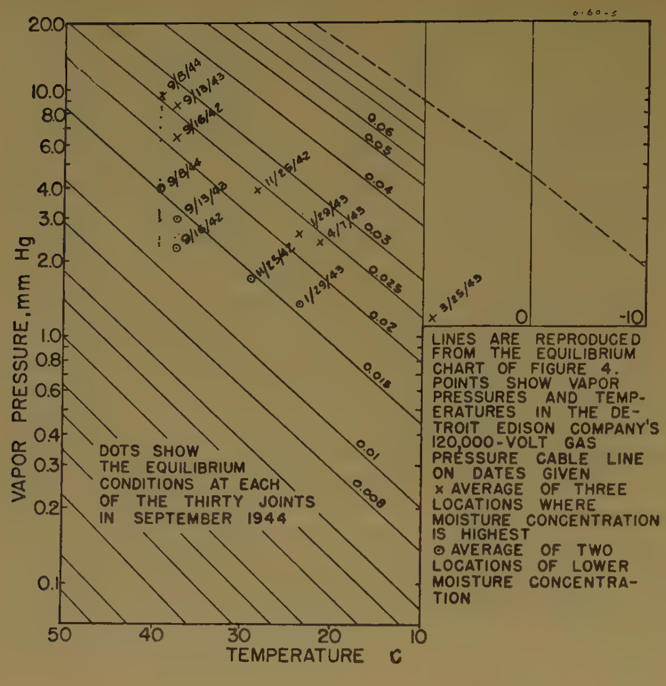


Figure 5. Similarity in equilibrium behavior between impregnated and unimpregnated cotton

1.7 per cent. The dew point of the gas space corresponding to a point that represents a given moisture concentration and selected temperature of the fibrous material is obtained by drawing a horizontal line from that point to the curve for water as shown by the arrows in Figure 6, and obtaining the dew point temperature as given by the point of intersection. Thus the chart shows that the dew point in the vicinity of tapes having the maximum moisture content of 5.0 per cent would be 26 degrees centigrade when the summer temperature of the cable line averages 38 degrees centigrade, and nine degrees centigrade when the winter temperature of the line averages 20 degrees centigrade. Thus it is shown that the drying treatments reduced both the maximum and the average dew points by approximately 15 degrees centigrade over the entire operating temperature range of the line, as shown.

Example 2

What concentration of water would a cotton protective tape acquire if the cable on which the tape is wrapped stands, until equilibrium takes place, in a room air conditioned to a relative humidity of 20 per cent at 30 degrees centigrade, and the cable thereafter is sheathed?

A 20-per-cent relative humidity at 30 degrees centigrade is equivalent to a moisture content of 6.0 grams per cubic meter which corresponds to a vapor pressure of 5.7* millimeters of mercury. A point representing 5.7 millimeters and 30 degrees centigrade on the chart indicates a moisture content of 2.6 per cent based upon the weight of dry unimpregnated cotton.

* Because the concentration of water in grams per cubic meter and the vapor pressure are nearly the same numerically up to 50 degrees centigrade, only a small error is introduced by multiplying the vapor pressure shown on the chart by the relative humidities to obtain the equilibrium vapor pressures. In this case, instead of 5.7 millimeters, a value of 20 per cent of 32 or 6.4 millimeters would be obtained.

Example 3

What concentration of water would preimpregnated kraft paper tapes acquire if they were exposed in the air-conditioned room described in example 2 long enough for equilibrium to be reached?

The point representing 507 millimeters of mercury and 30 degrees centigrade on the chart indicates a moisture concentration of 4.4 per cent in kraft paper.

The foregoing example does not imply that the moisture content of a preimpregnated paper insulation is high just because the equilibrium moisture content is high for the humidities that feasibly may be maintained in the taping room. Under conditions of normal operation the time of exposure is too short to allow significant absorption. For example, the average moisture content of the insulation of the cable previously mentioned was not 4.4 per cent but only 0.055 per cent, which is essentially the same as that for other well-made cables impregnated in a conventional manner. An equilibrium value that low would require, according to the chart, the relative humidity at 30 degrees centigrade to be of the order of 0.017 per cent. Obviously, the low rate of moisture absorption is of much more importance to the manufacturer of this type of cable than is the equilibrium concentration.

The two examples that follow illustrate the use of the equilibrium chart for problems involving condensation in transformers. For given equilibrium moisture content in the fibrous material, most of which is of wood origin, the probability of condensation is greater than for gas pressure cable. The reason is that the top of the transformer casing may be only a few degrees above outdoor temperature while the fibrous material is at the oil temperature, which may reach a daily maximum of 50 to 60 degrees centigrade. The equilibrium for transformers is more complicated than for the paraffin-impregnated tapes because the amount of water in the large mass of oil is not negligible as compared with that in the much smaller mass of fibrous materials. It is probable that a definite relationship also can be established for the equilibrium between the moisture content of oil and that of the gas space. Until such information is available, however, it is not possible to estimate the water content of the oil from dew point data. In the following the oil is considered only as a barrier that markedly reduces the rate at which equilibrium becomes established between the moisture contents of the gas and fiber.

Example 4

The gas in a new transformer had a dew point of ten degrees centigrade when operating at an oil temperature of 50 degrees centigrade while the outdoor temperature was well above the dew point of the gas. What was the probable moisture content of the kraft paper insulation within it?

A dew point of ten degrees centigrade corresponds to a vapor pressure of 9.2 millimeters of mercury, which at 50 degrees centigrade is shown by the chart to be in equilibrium with a moisture content of approximately 2.7 per cent in the insulation.

Example 5

The gas in a reconverted oil-conservator type of transformer had a dew point of 20 degrees centigrade when operating at an oil temperature of 42 degrees centigrade while the outdoor temperature was well above the dew point of the gas. How much water would have to be removed from the fibrous insulation of the transformer in order for the gas to have a dew point of minus ten degrees centigrade while the oil temperature was 50 degrees centigrade?

A dew point of 20 degrees centigrade, or a vapor pressure of 17.5 millimeters of mercury, indicates an equilibrium moisture concentration of 5.4 per cent at 42 degrees centigrade, according to the chart. A dew point of minus ten degrees centigrade or 1.95 millimeters of mercury indicates a concentration of 1.05 at 50 degrees centigrade. Thus it would be necessary to reduce the water content by approximately 4.35 per cent based upon the weight of the paper fibers. For a transformer containing 3,000 pounds of such insulation, it therefore

would be necessary to remove approximately 130 pounds or 15.6 gallons of water which does not include the amount that would have to be removed from the oil to reach the same equilibrium conditions.

Vapor Barriers

Consideration of the information given by the equilibrium chart leads to the conclusion that, if no vapor barrier existed between the cotton and the kraft paper tapes of the cable previously mentioned, equilibrium ultimately should become established between the moisture content of the cotton and kraft tapes with resultant reduction in the dew point of the gas space. The two kinds of tapes are present in the fiber-weight ratios of 22 to 485. Considering the region in which the final concentration of moisture in the cotton tapes was highest (2.65 per cent), if equilibrium were established with the kraft types, which had an average concentration of 0.055 per cent, the final moisture concentration of the cotton tapes would be 0.1 per cent and of the kraft tapes 0.17 per cent. The equilibrium vapor pressure at a temperature of 38 degrees centigrade then would be, according to the chart, approximately 0.05 millimeter of mercury, and the corresponding dew point approximately minus 45 degrees centigrade. Although establishment of the equilibrium between the moisture contents of the cotton and kraft tapes would be expected to be slow, there was no evidence that any exchange had taken place in over $2^1/_2$ years of operation. This is remarkable when it is realized that the conditions are the same as though the cable had been exposed continuously for over two years to a relative humidity of approximately 30 per cent. Furthermore, exposure for several months of unsheathed samples of the cable to Detroit's humid summer atmosphere out of direct contact with rain resulted in no apparent increase in moisture content of the kraft insulation. Of three samples of cable so exposed before the line was completed, the moisture concentrations were 0.01, 0.01, and 0.05 per cent, respectively, all of which values happened to be less than the average for unexposed samples. The author's interpretation of the reason the insulation does not become moist is that the metallized-paper shielding tapes bar the diffusion of water vapor into the insulation. This fact is believed to be of further practical significance, for no longer can it be said that plastic cable sheaths cannot be used for oil-impregnated cables on the basis that all such sheaths, although they provide a barrier for liquid water, are too permeable to water vapor. It seems entirely possible that if an elastomer can be developed to have suitable life in contact with oil on one side and ground water and air on the other, it should be a suitable sheathing for cables provided the outside of the insulation of each conductor is shielded by metallized paper or thin metal tapes. The latter without some sort of sheath are not adequate because they will not keep out liquid water nor provide mechanical protection. Use of a corrosion and abrasion-resistant elastic material that can be applied competitively with lead seems worthy of investigation, particularly for solid type cables. The elaborate "sandwich" scheme of Rihl and Heering[11] to use a plastic sheath seems unnecessary.

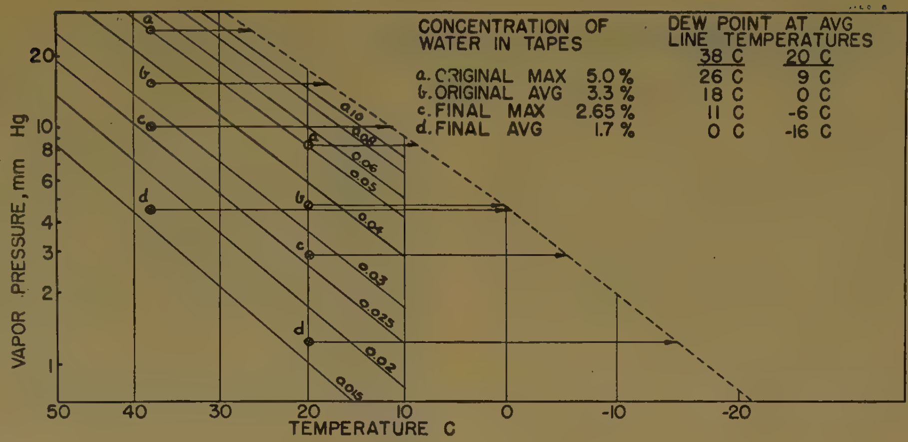


Figure 6. Use of the equilibrium chart in estimating the reduction in dew point that was effected by reducing the water concentration in impregnated cotton tapes

Conclusions

An equilibrium chart has been prepared from data from the literature to relate the equilibrium concentration of water in cotton or kraft paper with the vapor pressure of water, over a temperature range of 10 to 110 degrees centigrade. The concentrations of moisture covered by the chart range from 0.01 to 10 per cent for cotton and 1.7 times those values for kraft paper.

Although the equilibrium chart was prepared from data for unimpregnated fibers, data taken on The Detroit Edison Company's 120,000-volt gas-pressure cable line show that the chart is also valid for impregnated materials, at least for cotton impregnated with paraffin.

Examples are given of the use of the chart in predicting the equilibrium moisture content that fibrous materials will acquire when exposed to atmospheres of various relative humidities and temperatures, in predicting the dew points of the gas in enclosed electric apparatus, and in employing dew point measurements as a nondestructive analytical tool for determining the moisture content of fibrous materials in electric equipment.

Metallized-paper electrical shielding tapes placed over fibrous materials are shown to be remarkably effective in preventing diffusion of water into these materials. The significance of this phenomenon in connection with the use of an elastomer as a sheathing material, particularly for solid-type cables, is pointed out.

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A Multichannel Microwave Radio Relay System

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Synopsis: An 8-channel microwave relay system is described. Known to the Army and Navy as *AN/TRC*-6, the system uses radio frequencies approaching 5,000 megacycles. At these frequencies, there is a complete absence of static and most man-made interference. The waves are concentrated into a sharp beam and do not travel along the earth much beyond seeing distances. Other systems using the same frequencies can be operated in the near vicinity. The transmitter power is only one four-millionth as great as would be required with nondirectional antennas. The distance between sets is limited but by using intermediate repeaters communications are extended readily to longer distances. Short pulses of microwave power carry the intelligence of the eight messages utilizing *pulse position modulation* to modulate the pulses and *time division* to multiplex the channels. The eight message circuits which each *AN/TRC*-6 system provides are high-grade telephone circuits and can be used for signaling, dialing, facsimile, picture transmission, or multichannel voice frequency telegraph. Two-way voice transmission over radio links totaling 1,600 miles, and one-way over 3,200 miles have been accomplished successfully in demonstrations.

THE *AN/TRC*-6 is a combined transmitter and receiver designed specifically for radio relay purposes and includes multiplex facilities for providing eight two-way high-grade message circuits between points separated by an unobstructed optical path. Acknowledgment is made of the impetus and assistance given in this development by those concerned in a somewhat similar British development (British Wireless Set Numbet 10) conducted by the Ministry of Supply on behalf of the British Army.

In military use, one *AN/TRC*-6 set usually is transported by truck, and can be set up and placed in operation on a favorable site in a few hours. Figure 1 depicts an arrangement used by the Army in tactical movements, with the operational units housed in a truck. All of the microwave equipment is atop the 50-foot tower. Simultaneous two-way communication is obtained by the use of different radio frequencies for the two directions of transmission. Using separate antenna systems, the microwave transmitter and receiver-converter are mounted directly behind their respective 5-foot parabolic reflectors.

A sharply focused and highly directive microwave beam whose frequency is nearly 5,000 megacycles is used to carry the intelligence of the eight messages. This extremely high frequency affords communication channels free from static and most man-made interference. By the use of beamed radiation and reception, the transmitter power is only one four-millionth as great as would be required with nondirective antennas. Sharply beamed transmission combined with the absence of external noise means a small amount of transmitter power is adequate for communication over optical paths of considerable length. A peak power of a few watts serves for jumps as great as 100 miles. This power is obtained from a small tube no larger than an ordinary radio receiving tube, operating as a reflex velocity modulated oscillator ("reflex klystron").

Because of the sharp beam, because of the type of modulation used, and because transmission must be over an unobstructed optical path other sets using the same frequency can be operated in the near vicinity.

The sets were designed to be used in pairs to form a radio repeater, and it is practical to operate a considerable number of radio links in tandem. Because of the line-of-sight over which the system operates, the distance between sets is limited by the curvature of the earth, but by using intermediate sets as repeaters, communications are extended readily to hundreds and even thousands of miles.

When the set is used for military purposes, essential units are supplied in duplicate, as indicated by Figure 2 which is a close-up of the operational units. The portable test oscilloscope is for conveni-

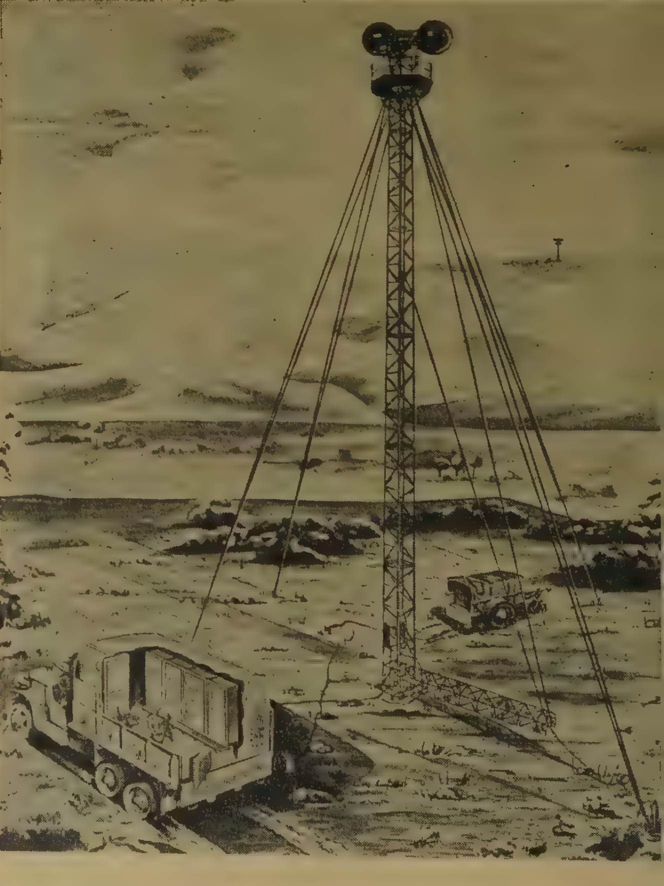

Figure 1. *AN/TRC*-6 with operational units in a truck

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Figure 2. *AN/TRC-6* set up as a field terminal

ence in monitoring. A high voltage rectifier for the transmitter is located at the bottom of the tower.

An optional and improved arrangement representing a recent development and known as *AN/TRC*-6 (*XC*-4) is now available. This improved equipment features one antenna instead of two, a lighter weight tower presenting a less conspicuous target, and the location of the microwave transmitter and receiver-converter on the ground instead of atop the tower. The radio transmitter and receiver-converter connect through wave guide filters to a single antenna system also located on the ground. The parabolic ground reflector (Figure 3) is beamed on a wire mesh 48-degree plane reflector atop the lightweight 50-foot mast. The extra loss introduced by the single antenna system is about three decibels.

Unlike conventional communication equipment which transmits a continuously modulated wave, *AN/TRC*-6 makes use of pulse modulation. In such a system, the transmitter is caused to emit short pulses of microwave power. These microwave pulses are substantially constant in amplitude and frequency. Eight 1-microsecond pulses (Figure 4), one for each channel, are transmitted in sequence and are preceded by a 4-microsecond synchronizing pulse called a marker. This sequence of marker plus eight channel pulses repeats itself 8,000 times a second and is called a frame.

The intelligence of each channel is conveyed by varying the exact position in time of the channel pulses. The phrase pulse position modulation has been applied to this method. The eight channels share the operating time of the transmitter, each using the radio path in turn. Such a sharing process utilizes a multiplexing principle that for many years has been known as time division. Pulse modulation, therefore, not only provides for position modulating the pulses but also permits multiplexing the channels by time division. No detectable distortion of the recreated message in the final receiver need be inherent in this method of transmission, nor is the scheme limited to eight channels. Systems using these same principles could be designed readily for many more channels.

In carrying out the pulse modulation method, the voice wave of each incoming message is sampled 8,000 times a second. Each sampling subsequently results in a pulse of microwave power leaving the transmitting antenna. With no voice input, the pulse associated with a particular channel recurs 8,000 times a second or every 125 microseconds. When the channel is busy, the pulse occurs earlier or later, depending on whether the voice wave is positive or negative. The exact displacement of a pulse from its unmodulated position is linearly proportional to the magnitude of the voice wave at the time of sampling.

The modulation of channel 2 in this manner is illustrated in Figure 5 for the case of a particular sinusoidal input of 1,000 cycles. When the input signal is zero, namely, at points A, E, and I in the figure, the channel pulse will occur at the middle of the position allocated to the channel. Successive diagrams depict the change in the relative position in time of the channel-2 pulses from frame to frame for nine successive frames.

The eight voice circuits connect to the common unit (Figure 6) and go to the transmitting portion of the multiplex which generates marker and channel pulses. These pulses are cabled to the transmitter and are used to turn the transmitter (transmitter-oscillator) on or off for intervals of time corresponding to the length of each pulse. Thus, d-c or video pulses from the transmitting multiplex are translated into pulses of microwave power. These microwave pulses are conducted to the focal point of the transmitting paraboloid where they radiate to illuminate the reflector, the pattern of the reflected radiation being sharply beamed.

Attenuated microwave pulses from the distant transmitter are picked up by the receiving parabolic reflector (Figure 6) and go to the receiver (receiver-converter). The converter changes the microwave pulses to *IF* (intermediate frequency) pulses centered about 58.5 megacycles. After amplification, the *IF* pulses go to the remainder of the radio receiver located on the ground.

Here the *IF* pulses are amplified further, detected; and the resulting video pulses fed to an amplifier. The receiver has automatic tuning so that it is at all

Figure 3. *AN/TRC-6* featuring two-way transmission with a single antenna

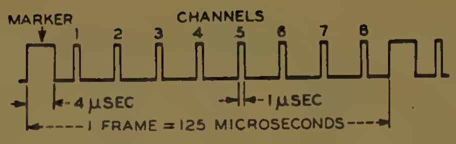


Figure 4. Pulse array at output of transmitting multiplex or input to receiving multiplex

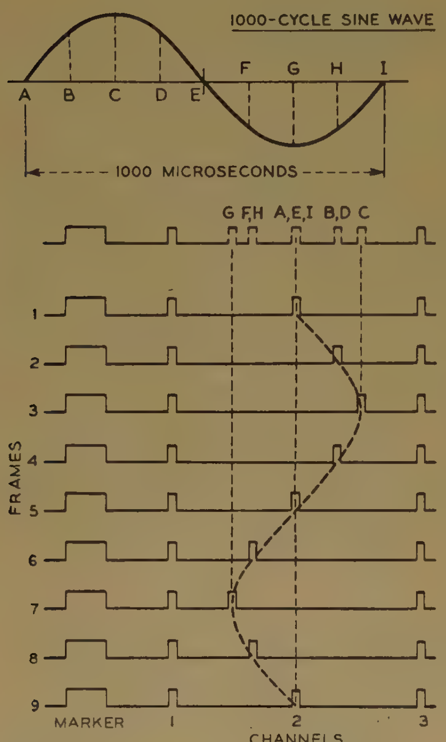


Figure 5. Modulation of channel-2 pulses

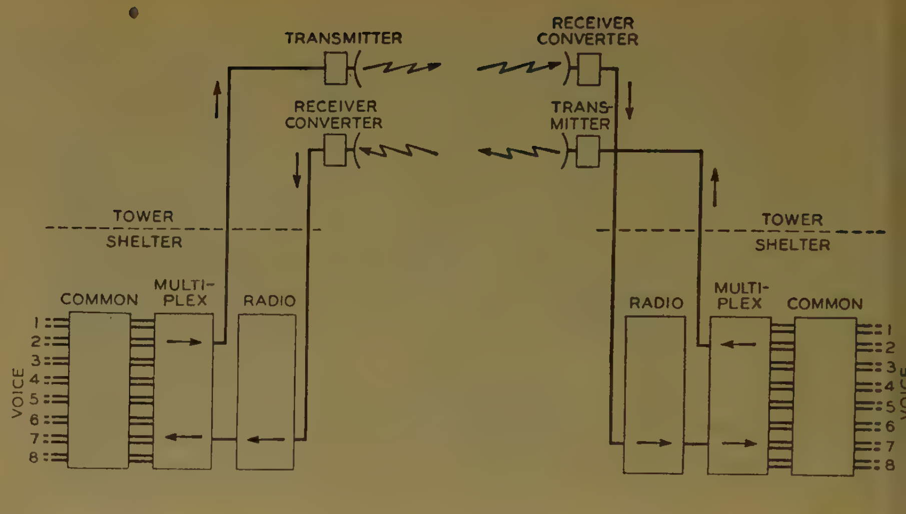


Figure 6. Block diagram of single link system

times correctly tuned to the distant transmitter. The receiver also has automatic volume control so that the output pulses are substantially constant despite fading.

The receiver has a special feature termed a slicer or double clipper whose function is to carve out a narrow slice (about 5 per cent) from each received pulse. The slicer (Figure 7) substantially frees the pulses of noise.

The output of the slicer goes to the receiving multiplex. Here the marker pulse is identified because of its greater length and causes the receiving multiplex to operate "start–stop."

The sequence of events triggered off by recognition of the marker opens, successively, electronic "gates" to each of the eight channels in turn. After a particular channel gate is opened, the next pulse passes through and is caused to trigger another circuit which starts a direct current flowing. This flow of direct current continues until stopped by the closing of the gate. The starting and stopping of a direct current in this manner generates a length modulated pulse whose trailing edge is fixed but whose leading edge depends upon the exact timing of the received channel pulse. This represents the conversion of pulse position modulation to pulse length modulation. The voice signal is derived by passing the length modulated pulses through a low-pass filter.

Specific Embodiment

Method of Operation of the Multiplex

Figure 8 is a block diagram of the transmitting section of the multiplex. An 8,000-cycle oscillator fixes the frame frequency and produces a rectangular wave for starting the marker pulse generator and eight pulse position modulators.

The marker pulses are generated by differentiating the oscillator output and applying the resulting pulses to a tube so operated as to be insensitive to positive pulses. Each negative pulse produces an output pulse which, when limited or clipped, results in a nearly rectangular 4-microsecond pulse at the start of each frame.

Except for channel 1, which will be described separately, short positive pulses are required for starting the pulse position modulators. For channels 7 and 8 it is possible to derive these directly from the oscillator by a differentiating circuit. Exciter circuits, similar in design to the marker generator, start channels 2 to 6. Use of separate exciters for odd and even numbered channels reduces interchannel crossstalk.

Except for channel 1, each position modulator is a multivibrator. This multivibrator is a 2-stage resistance-capacitance coupled amplifier with its output fed back to the input. The circuit is so operated that the first stage normally is cut off and the second conducts. Application of a positive trigger pulse to the grid of the first stage causes it to conduct and cut off the second stage. After an interval of time, depending mainly upon the time constant of the *R-C* circuit which connects the two stages, the multivibrator relaxes, that is returns to its original condition with the first stage cut off and the second conducting. Each trigger pulse results in an output pulse, the duration of which is determined primarily by the time constant. The time constants are adjusted so that each multivibrator relaxes at the time assigned to that particular unmodulated channel pulse.

Although the time of relaxation of the multivibrator is controlled primarily by its time constant, this time also can be altered by superposing a varying voltage on the grid of the second stage. This grid is the point at which voice frequency voltage is applied from the voice amplifier of that particular channel. Consequently, the trailing edges of the pulses developed by the multivibrator are position modulated by the voice input. A 1-microsecond pulse, produced in a generator common to four channels,

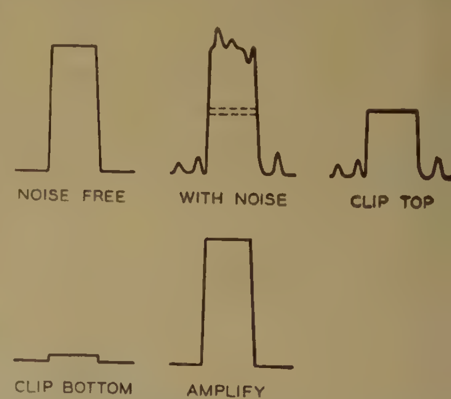


Figure 7. Reduction of noise by slicer or double clipper

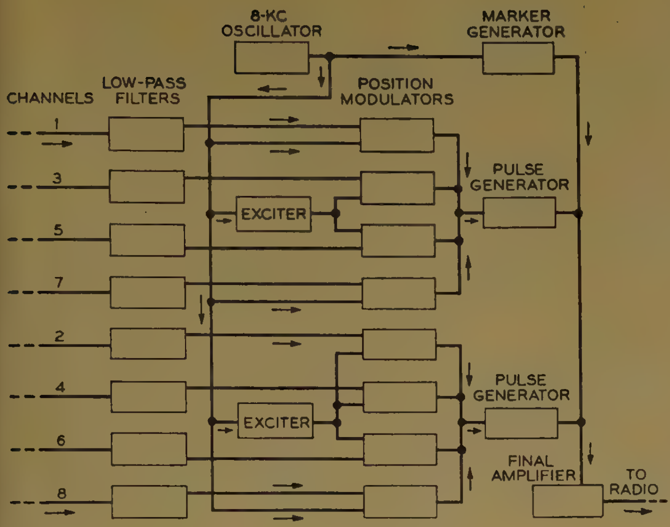


Figure 8. Block diagram of transmitting multiplex

is derived from this trailing edge. The two groups of channel pulses and the marker are combined and go to the radio transmitter.

Because little time exists between the marker and the first channel pulse, channel 1 differs from the other channels in that it does not require a multivibrator to delay the channel pulses. For this channel, therefore, a simpler circuit is possible. This circuit produces a length modulated pulse by clipping a saw-tooth shaped pulse at a voltage determined by the voice sample. The desired channel-1 pulses are derived from the leading edges of these length modulated pulses.

In the receiving section of the multiplex (Figure 9), video pulses from the radio are amplified and the circuit branches into two paths, one of which serves to select the marker pulse. The marker selector is so arranged that its output voltage is proportional to the duration of an input pulse. Marker pulses produce voltage peaks about four times those resulting from channel pulses. The marker amplifier is negatively biased so that the channel pulses do not cause plate current to flow. Therefore only the marker is passed. A new pulse derived from the trailing edge of the selected marker is fed to a multivibrator (called square wave generator) which thus is synchronized accurately with the marker. This symmetrical square wave is used either directly or through intervening sweep generators to operate the gate generating circuits.

To separate the channels, properly timed gate pulses are used. Each gate is a rectangular pulse 13 microseconds long. The gates must be delayed by varying amounts from the edges of the square wave. This is accomplished by generating sweep or saw-tooth voltages and starting the gate pulses at the time the respective sweep voltages cross a reference voltage. The circuit constants for the various channels are different so as to vary the rate of rise of the sweep voltage and thereby locate each gate at its assigned time.

It was noted previously that the received video pulses branch into two paths, one of which selects the marker. The second path includes a 2-stage amplifier which produces new pulses derived from the trailing edges of the received pulses.

These new pulses are fed to the input of the eight channel converters. To each converter is fed also the output of the appropriate gate generator. The gate pulse voltage plus its corresponding channel pulse voltage is appreciably greater than the voltage of the other channel pulses. For example, gate 1 voltage plus channel 1 voltage is greater than channel pulses 2 to 8. Channel 1 is separated from the other channels and this position modulated pulse is converted to a length modulated pulse by a multivibrator so operated that it can be triggered by the combined gate and channel pulse voltages but not by either alone. This circuit not only separates the channels but also converts pulse position modulation to pulse length modulation. When the channel pulse is superposed on the gate pulse, sufficient voltage is obtained to start the multivibrator so that the first stage conducts and remains conducting for the remainder of the gate pulse interval. At the end of the gate pulse, the first stage in the multivibrator circuit is cut off, thereby returning the multivibrator to its original condition. Because the multivibrator is started at a variable time dependent upon the position of the channel pulse and is cut off at a fixed time established by the end of the gate pulse, the output pulses vary in length in accordance with the instantaneous position of the input pulses. By passing these length modulated pulses through a low-pass filter, the audio signal is obtained and all modulation products above the highest audio frequency are eliminated.

Special Radio Features

While two radio frequencies suffice for operation over a single link, a choice of four frequencies is available as a further means of reducing interference between systems or between sets of the same system when repeaters are operated in tandem. Although designed to operate on fixed frequencies, the equipment is capable of being adjusted continuously to operate at any frequency between the extremes. Both transmitter and receiver-converter use small oscillator tubes of the reflex velocity-modulated type

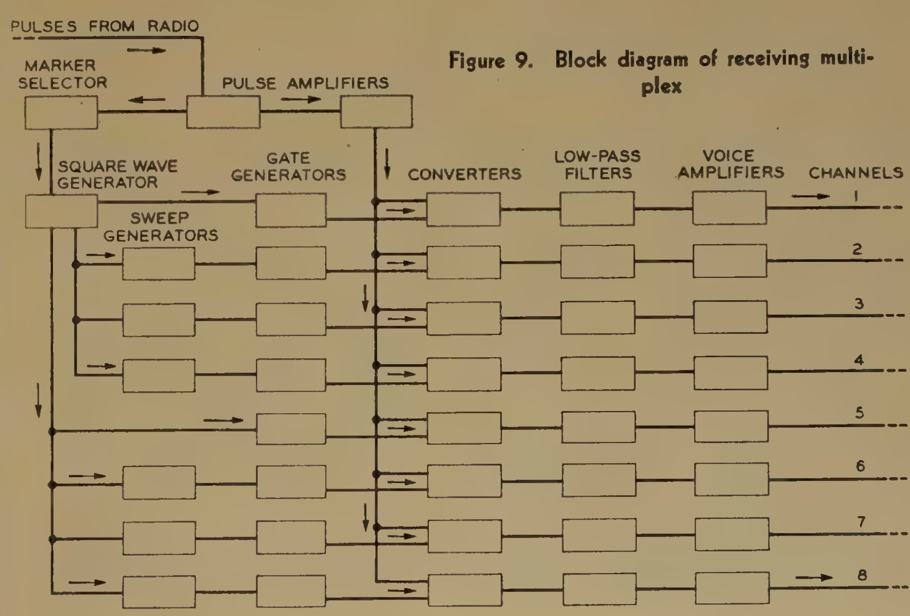


Figure 9. Block diagram of receiving multiplex

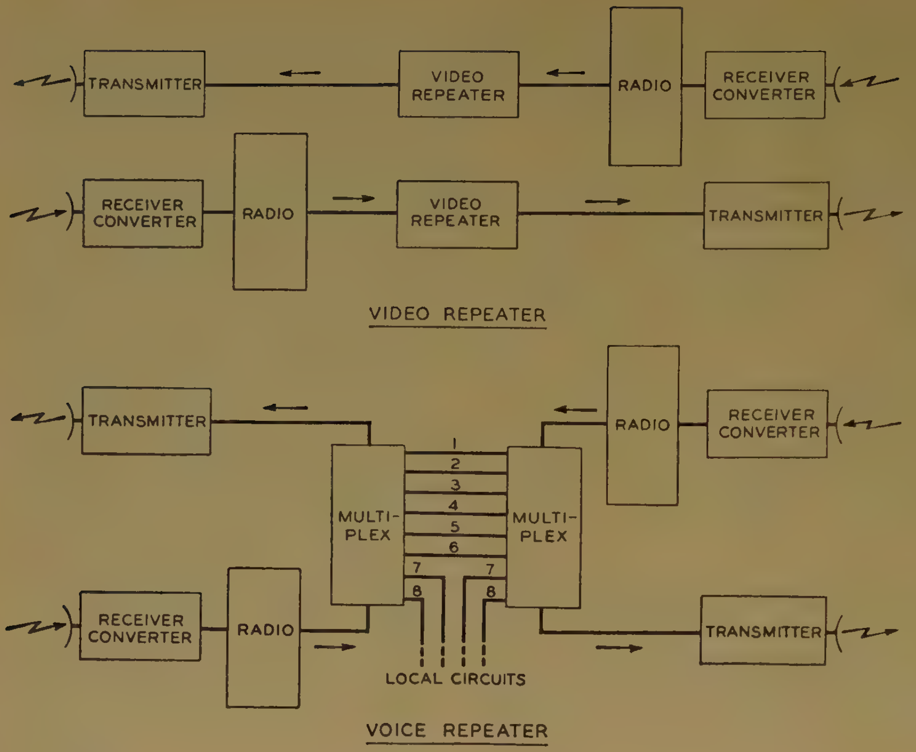


Figure 10. Alternative repeater arrangements

having a flexible cavity which is mechanically varied to get the approximate frequency. The exact frequency is obtained by varying the repeller potential.

Each antenna has a gain of approximately 33 decibels over a simple half-wave antenna. The polarization is normally horizontal but vertical polarization can be used if desired.

A feature of a pulse modulated oscillator is that because power is not generated continuously, the heating of the tube is less than for a continuous wave system and advantage may be taken of this by increasing the applied voltages so as to get increased power during the pulsing intervals.

The radio transmitter is a simple self-excited oscillator which oscillates or not depending on whether a pulse is present. This oscillator is pulsed by a modulator tube whose plate connects to the cathode of the oscillator. The repeller electrode is pulsed simultaneously with a fixed fraction of the voltage applied to the cathode in order to reduce frequency modulation to a minimum. A peak pulse power of 2 to 10 watts is obtained depending upon the type of tube and frequency.

Radio frequency power from the oscillator passes through a short coaxial lead which radiates into a wave guide through an impedance matching arrangement. The wave guide is tuned with a plunger at one end and delivers power to the antenna at the other end. A coupling arrangement is provided on the latest model to permit in-service measurement of microwave power without affecting the power delivered to the antenna. Microwave power is measured with a portable direct-reading power meter.

In the latest model of the receiver-converter, the beating oscillator is fed to a wave guide through an attenuator and coupling arrangement which permits amplitude adjustments or frequency measurements without regard to the impedance presented to the beating oscillator by the antenna.

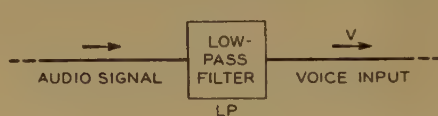


Figure 11

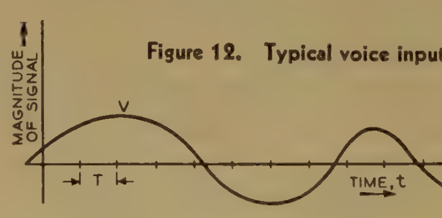


Figure 12. Typical voice input

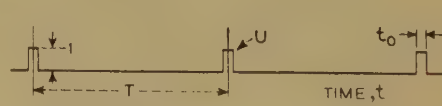


Figure 13. Unit sampling function, *U*

$$U=k+2k\sum_{m=1}^{\infty} A_m \cos mCt$$

$$f_c=\frac{C}{2\pi} \qquad \frac{1}{f_c}=T$$

$$k=\frac{t_0}{T} \qquad A_m=\frac{\sin mk\pi}{mk\pi}$$

Video Repeater

The two radio sets forming a repeater can be joined either at voice or video frequencies (Figure 10). Joining at voice frequencies is a convenient arrangement if message circuits are dropped or added.

When sets are joined at video frequencies, there is a saving in equipment because the multiplex boxes are not required. Instead, a video repeater is used. Located in the common box, this repeater provides order wire facilities for monitoring and modulating channel 1 so that a relay attendant will be able to signal and talk to any other station along the route. In addition, the received pulses are lengthened to compensate for their having been shortened, chiefly by the preceding transmitter.

Special Equipment Features

The mechanical and equipment features of *AN/TRC-6* were governed largely by Signal Corps specifications. Important objectives such as ruggedness, portability, maintenance, storm-proofing, ease of setup, weight, and dependability were given proper balance to obtain a set suitable for military use.

The 50-foot sectionalized welded aluminum tubing tower consists of six 8-foot sections from two to three feet square with provision for omitting sections to obtain lesser heights. The four intermediate sections are tapered and may be telescoped together for shipping. The top section rests in the base section together with the hoisting boom during shipment.

With the tower properly guyed, the movement of the antenna beam pattern is restricted to a fraction of a degree even in a 60-mile wind. The sets have oper-

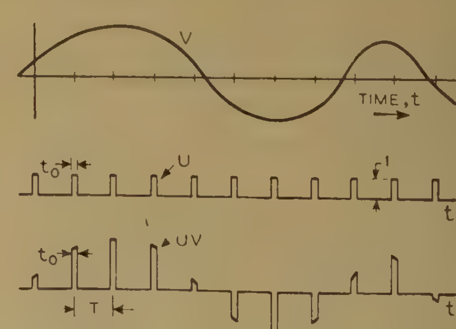


Figure 14. Diagram of *V* and *U*

UV is the result of sampling *V*

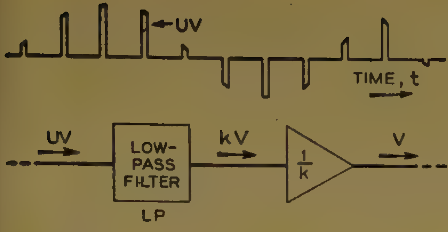


Figure 15. Passing *UV* through a low-pass filter and amplifier to obtain *V*

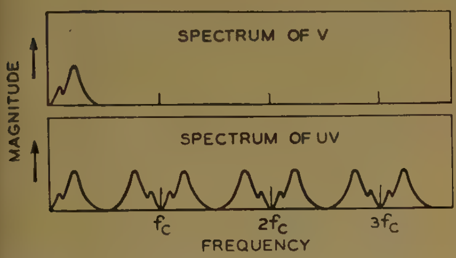


Figure 16. Spectrum analysis of *V* and *UV*

ated successfully through two hurricanes. The electrical and mechanical performance have proved satisfactory in actual field usage even when the antennas, top beam section, platform, tower, and guys were coated with more than three-quarters of an inch of ice.

Pulse Modulation Principles

The discussion to follow is a treatment of pulse modulation including the sampling principle and time division. To enumerate important features:

1. It is the sampling principle which specifies the least number of points required to reproduce a changing voice wave.

2. It is the job of the modulated pulses to carry this information to the distant receiver.

3. Pulse position modulation, like frequency modulation, reduces noise at the expense of band width.

4. A method of scaling is described whereby time division is regarded as a scale model of a single channel system.

In carrying out the pulse modulation method, the voice signal usually is passed first through a low-pass filter (Figure 11) to cut out all frequencies above an upper limit which is less than one-half the sampling frequency, f_c. After filtering, the signal is designated V and termed voice input.

V (Figure 12) is the instantaneous value of a short portion of a voice signal which might have passed the low-pass filter of Figure 11. The sampling principle teaches that if V is sampled instantaneously at regular intervals T, where T is less than a half period of the highest frequency in V, then the samples contain all of the information of the original signal. The magnitude of a sample is the information a modulated pulse is required to carry to the distant receiver.

The unit sampling wave (Figure 13) is designated U and will be used presently to sample the voice. Mathematically U is regarded as equal to a d-c term plus harmonics of the sampling frequency, f_c. The interval between pulses is designated T and equals $1/f_c$. The ratio of pulse length to the interval between pulses is designated k.

The voice input and unit sampling wave are repeated at the top of Figure 14. Because U is either one or zero, the product UV is an analytical process for sampling the voice. The result is a series of positive and negative pulses. When U is one, the product is equal to the voice input and at all other times zero.

Because UV is similar in appearance to amplitude modulated pulses (Figure 15), it is not surprising to find that an attenuated replica of V is obtained simply by passing UV through a low-pass filter. In the process of sampling and filtering, V is reduced by a factor k. Amplifying by a factor $1/k$ restores V to its original value.

To bring out more concretely that passing UV through a low-pass filter produces the original signal, Figure 16 presents a spectrum analysis of V and also UV. The upper diagram is the spectrum of V. After sampling V becomes UV. The spectrum of UV is the spectrum of V, small but exact, together with upper and lower sidebands about harmonics of f_c.

Remembering UV is the result of sampling, we can draw a conclusion; if, in a communication system, the resultant sampling UV is transmitted to the distant receiver either by direct transmission or by conversion to pulse position modulation, the voice signal can be derived simply by passing through a low-pass filter followed by an appropriate amplifier.

The simple physical concepts of pulse position modulation (Figure 17) will be reviewed by considering an idealized 1-channel system. With no voice input, pulses of fixed amplitude and length recur at regular intervals T. When the channel is busy, each pulse is displaced from its unmodulated position and occurs earlier or later depending on whether the voice signal is positive or negative. The exact displacement is linearly proportional to the instantaneous value of the voice wave at the time of sampling. With full input, the maximum excursion, neglecting the length of the pulse, is plus or minus $T/2$ or $\pm 1/2\, f_c$.

Figure 7 illustrates the reduction of noise by the slicer or double clipper. A practically noise-free pulse is delivered by the transmitting multiplex. A long radio path attenuates the pulse and noise is introduced by the receiver, producing a typical pulse with noise as indicated. The double clipper works in two steps. First, the top of the wave is shaved off reducing noise superimposed on the pulse. Second, the bottom of the wave is chopped off leaving a nearly noise free but attenuated pulse. This is amplified

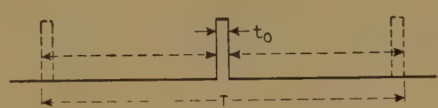


Figure 17. Single-channel pulse-position-modulation system

$$\pm\frac{T}{2} = \pm\frac{1}{2f_c}$$

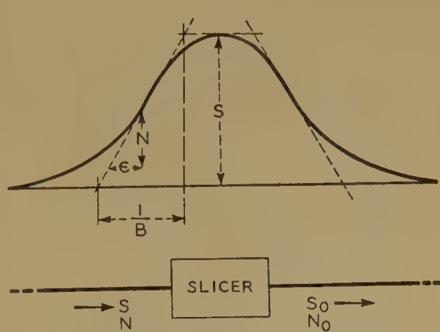


Figure 18. Equation for signal-to-noise improvement resulting from slicer

$$\frac{S}{N} = \frac{1}{B\epsilon} \qquad \frac{1}{\epsilon} = \frac{SB}{N}$$

$$\frac{S_0}{N_0} = \frac{1}{2\sqrt{2\epsilon f_c}} \qquad \frac{S_0}{N_0} = \frac{S}{N} \times \frac{B}{2\sqrt{2 f_c}}$$

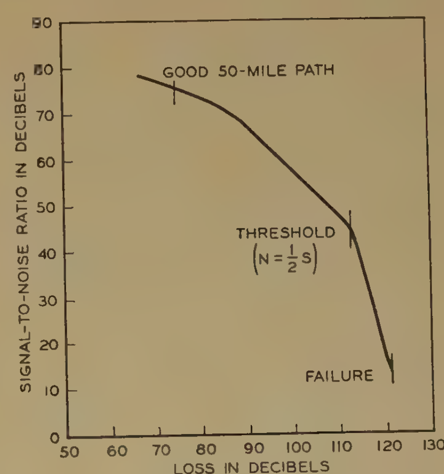


Figure 19. Decibel signal-to-noise ratio versus path loss

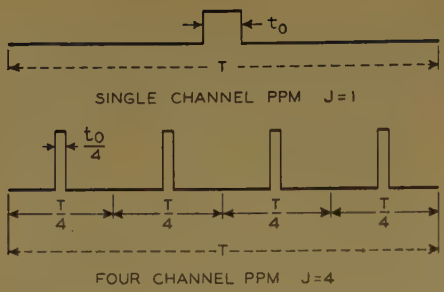


Figure 20. Formation of *J* channels by time division

and delivered to the receiving multiplex. It should be noted that whereas the effects of noise on amplitude can be practically eliminated by the slicer, nevertheless, the effects of noise on the exact timing of the pulse remain. However, the magnitude of this error in timing can be reduced by increasing the band width.

Because the exact time of occurrence of these nearly noise-free pulses is changed by the presence of noise at the input to the slicer, this effect, namely, the variation in exact time of occurrence of the pulses, will register as additional noise in the final receiver. This additional noise is evaluated by an approximate formula (Figure 18) for the signal-to-noise improvement due to the slicer (see Appendix). $B/2\sqrt{2f_c}$ is the improvement where B is the *IF* band.* This is the improvement for a single channel system whose maximum excursion approaches plus or minus $T/2$.

Because of the slicer, noise affecting the remainder of the system is restricted substantially to that which occurs during the time that the pulse amplitude is within the range of the slicer, provided the peak amplitude of the noise is less than one-half the peak amplitude of the signal. As the noise exceeds this limiting value or threshold, the ability of the receiver to distinguish between noise pulses and signal pulses begins to deteriorate (Figure 19). For values of peak noise below the threshold, the improvement in signal-to-noise ratio resulting from the slicer is constant.

A feature of pulse position modulation is that for fixed average power out of the transmitter, the peak power can be increased as the pulse length is reduced. Thus, if the band width is increased K-fold, it is assumed the pulse length is divided by K and that the pulse power is increased K-fold. Assuming the noise has characteristics similar to resistance noise, the signal-to-noise ratio at the input to the slicer and, therefore, the threshold, are independent of band width. Because of the K-fold wider band, the maximum rate of rise and fall of the pulse will be increased K-fold, and consequently the signal-to-noise ratio at the output of the slicer further is increased K-fold which corresponds to a further improvement in decibels of $20 \log_{10} K$. This is in contrast to a conventional frequency modulation system where, over a path of specified attenuation, the threshold represents a limitation on the signal-to-noise ratio obtainable with a given amount of carrier power.

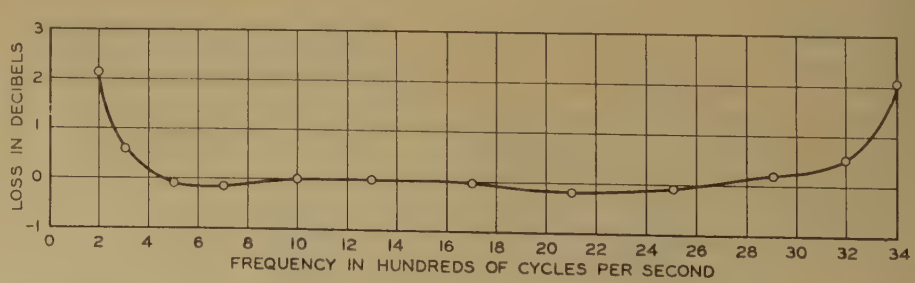
Passing to the theoretical concept of time division, a time division multiplex can be treated by first considering an idealized 1-channel system. Assuming one channel and maximum excursion of the pulse, the band width will depend upon the signal-to-noise improvement sought. For double sideband transmission, the *IF* band will be twice the video. The band width of the transmitter should equal that of the receiver. The pulse should be as short as the band width of the system will tolerate.

To provide J channels by time division (Figure 20), all significant band widths are increased J-fold; the length of each pulse is divided by J; the peak power of each pulse is multiplied by J; J pulses appear in each frame; the average power of the transmitter is increased J-fold; and the modulation or maximum excursion of each pulse is divided by J. Forming a scale model in this fashion, the J-channel system will have the same threshold and same signal-to-noise ratio as the idealized 1-channel system.

Transmission Performance

Microwave propagation at frequencies of 4,000 to 5,000 megacycles is subject to fading, and observed diurnal and seasonal effects are evidence of a close connection between microwave propagation and local meteorological conditions.

Figure 21. Typical transmission–frequency characteristic



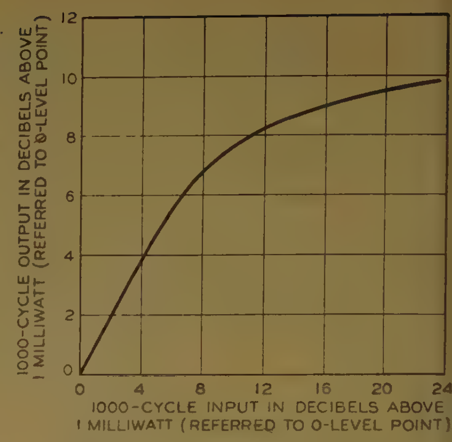


Figure 22. Typical load performance

However, the presence of rain or snow in the transmission path has given no evidence of causing serious increases in attenuation. Although the radio propagation may vary by very substantial amounts, the transmission stability of the message circuits is extremely good.

The eight message circuits which each *AN/TRC-6* system provides are high quality telephone circuits and meet commercial standards for long distance telephone transmission. The individual message circuits may be terminated 2-wire or 4-wire and may be used to transmit signaling, dialing, facsimile, pictures, or multichannel voice frequency telegraph. In the latter instance, a single message circuit will handle as many as 18 separate teletypewriter facilities.

As indicated by Figure 21, the transmission–frequency characteristic of each of the eight channels varies less than one decibel from 300 to 3,300 cycles, and, as shown by Figure 22, each channel is capable of handling inputs up to about plus eight decibels above a milliwatt at the zero level point (2-wire input) without appreciable overloading. The load capacity of the receiving multiplex is sufficient so that when terminated 4-wire, the circuit can be adjusted to nine decibels gain. Receiving potentiometers provide a continuous adjustment of channel gain over a 15-decibel range.

Inspection of Figure 23 shows that the circuit noise observed on a single link is dependent upon the attenuation in the

* An evaluation of the signal-to-noise ratio that is similar in principle is included in an article entitled "Pulse Time Modulation" by E. M. Deloraine and E. Labin in Electrical Communication, volume 22, 1944, number 2, pages 91–8.

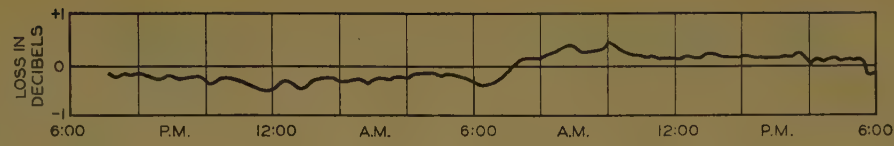
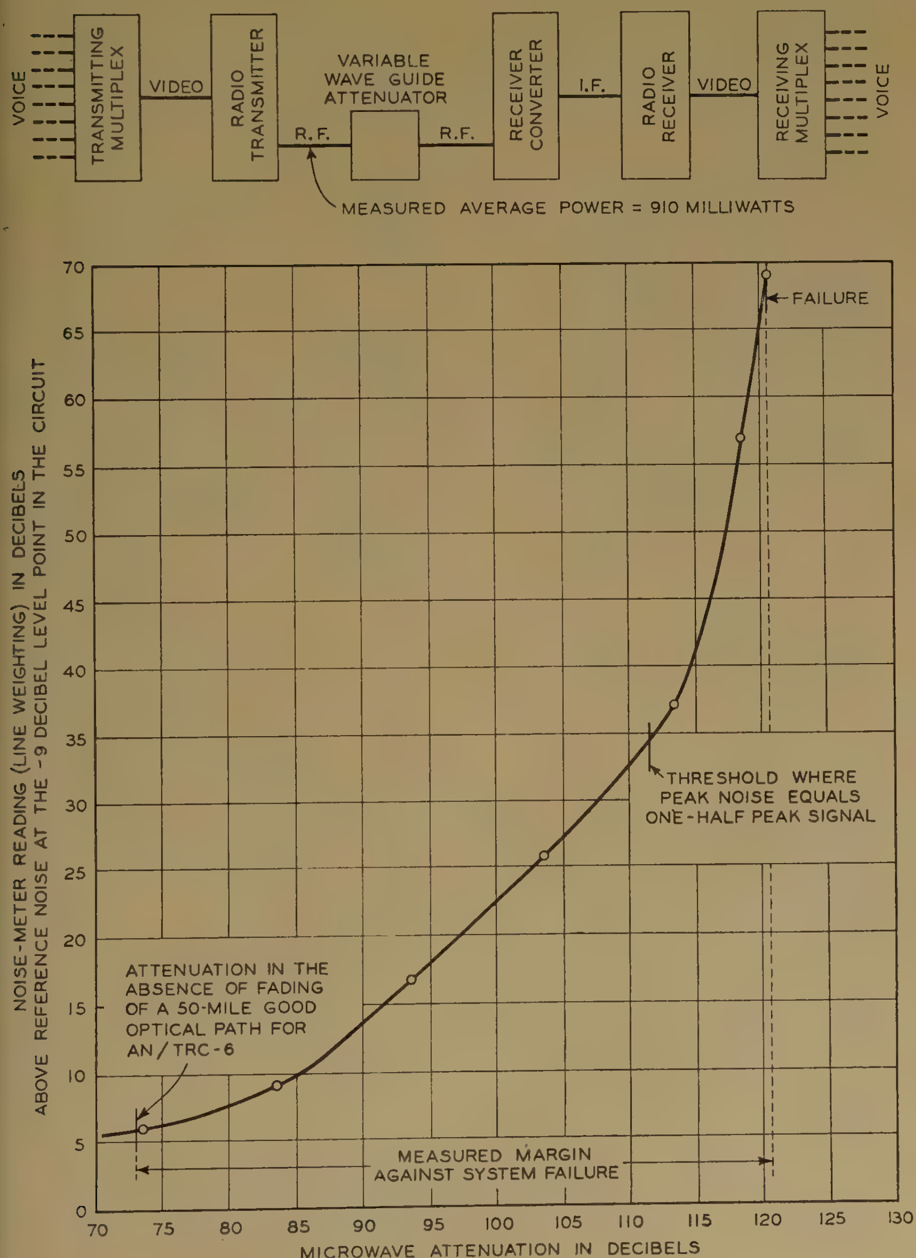


Figure 24. Transmission stability of 3,200-mile circuit

path. For a many link system, this noise increases with the number of repeaters. However, the individual message circuits are particularly quiet. For example, over a good 50-mile path, the noise in the absence of fading is approximately six decibels above reference noise when measured at a receiving level nine decibels below zero level. In terms of rms signal-to-noise ratio, assuming the signal is plus eight decibels above a milliwatt at zero level, this corresponds to a signal-to-noise ratio of 75 decibels. Further inspection of Figure 23 will show that on a 50-mile path, a margin against failure of approximately 48 decibels has been built into the set, and fades approaching this magnitude can be encountered before the system fails.

Figure 23. Message circuit noise for a single link versus path loss



The intervals of time that are significant in determining the exact position in time of a channel pulse are small. For example, in Figure 23, if the relative random jitter in the trailing edge of the channel pulses had been as great as one thousandth of a microsecond, the noise meter readings for sufficiently low values of path attenuation would have been slightly more than 3 decibels higher than the experimental values observed.

Two-way voice transmission over radio links totalling 1,600 miles and one-way over 3,200 miles have been accomplished successfully in demonstrations. To do this, ten radio sets were set up to form five two-way 8-channel systems each operating simultaneously over the same 40-mile air path. The five links were connected in tandem to form a two-way 8-channel system 200 miles long with four intermediate repeaters. By connecting all channels in tandem a two-way 1,600-mile circuit was obtained. Finally, by connecting both directions together, excellent 1-way transmission was obtained over an air path 3,200 miles long. To illustrate the over-all stability of the 3,200-mile circuit, Figure 24 is a typical 24-hour recording of the received level of a 1,000-cycle test tone and it can be seen that the maximum variation is less than plus or minus 0.5 decibel.

Conclusion

Although this equipment was designed for military use, its basic principles and design features can be used to provide telephone and other communications to the public. Sharply beamed radiation and reception, pulse modulation and time division multiplex are all factors which tend to make such systems attractive and their economic field in competition with other alternatives needs to be determined.

Appendix

Noise will modify the times at which the incoming pulses reach the slicing or triggering level. Figure 18 shows how a small noise voltage N, superimposed on a video pulse of amplitude S, changes the triggering time by an amount ϵ. From the geometry of the figure, $S/N = 1/B\epsilon$. After passing the IF amplifier of band width B and second detector, the approximate slope of the leading edge of the pulse at the slicing level is SB and the corresponding "time of rise" is $1/B$. When the output of the slicer is passed to the receiving multiplex, the error in timing ϵ will cause a noise voltage N_0 in the output of the receiving multiplex.

Assuming a 1-channel system (Figure 17) and supposing that the maximum excursion of a pulse can assume its limiting value of plus or minus $1/2f_c$, plus or minus $1/2f_c$

An Improved Azimuth Indicating System for Aircraft

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Synopsis: Because of the violent maneuvering possibilities in modern aircraft, the usefulness of azimuth indicators in avigation depends on circumstances not existing in marine navigational applications. New problems in direction indicating instruments therefore have been introduced, such as the perplexing problem of obtaining indication after loops or rolls. Azimuth indicating systems now used in aircraft have not given a completely satisfactory solution to such problems.

This paper summarizes the progress from the direct reading compasses to monitored gyroscopes and concentrates on the unique problems in aircraft where space limitations, reliability, and operation at all altitudes are important. It indicates how azimuth indicators and auto pilot controls close to the ideal are obtainable by the use of special features including a novel "nontumbling" mechanism and an electromagnetic type of gimbal adjusting device.

DURING the past few years more emphasis has been placed on the development of a directional indicating and control system which will meet the new rigid requirements for avigation as well as the older well-established needs of navigation which are still used in guiding aircraft. The first systems for use in aircraft followed the natural development trend towards improvement of navigational compasses. However, the rotational and translational accelerations which are experienced in modern aircraft bring forth new and, until recently, unsolved problems in azimuth indication. A major and perplexing problem is that of obtaining indications during steep banks, rolls, and loops. It is the purpose of this paper to point out the principal features needed in an azimuth system for use in modern aircraft and to indicate how some of the major problems can be solved. It will be shown how a modern measurement system has been designed which closely approaches the ideal.

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Existing Systems

Before stating the ideal characteristics, it is desirable to review briefly some of the systems utilized for indicating and controlling aircraft directions. The direct reading compass is simple, rugged, light in weight, and small in size. Under favorable conditions it is reliable and accurate. Because its indications often fluctuate rapidly and are influenced by proximity to magnetic fields or magnetic materials, its use is often limited to emergencies. It also displays the northerly turning errors which are well known to pilots and others familiar with compasses.[1] The remote indicating compass helps to solve the difficulty resulting from magnetic fields and materials by locating the direction sensitive element in a position where such influences are minimized. However, it does not solve the problems of northerly turning errors or pointer oscillation.

In certain cases the stabilization of the magnetic compass has been found desirable[2] in order to eliminate the effect of the vertical component of the earth's field which is responsible for the northerly turning errors. This is done in the gyrostabilized compass by mounting the direction sensitive element on a vertical axis gyroscope. The element thus is maintained in a level position during maneuvers and while flying in rough air. However, because the vertical gyroscope is kept vertical by gravity control, it will have some errors during maneuvers of long duration since centrifugal forces influence the apparent direction of the gravitational force. It also has the disadvantage of heavy weight and large size in the forms that are generally available. Although in this general design the compass transmitter including the gyroscope can be located remotely to minimize magnetic effects, its location must be selected carefully to provide sufficient space and to be sure that vibration will not damage the gyroscope bearings. Modern high speed airplanes allow only a few inches of height in the wing tips where it is often desirable to locate the transmitter.

Another class of azimuth indicator is the directional gyroscope which in its electric form[3] has been used mainly as an indicating and control instrument in a relief type of autopilot.[4] This device gives an indication that is well-stabilized but is subject to gradual drifting or wandering and must be reset frequently to the magnetic compass heading. A part of the drift may be the result of gimbal bearing friction, unbalance, or mechanical imperfections, but a more fundamental cause is the rotation of the earth which produces an apparent drift depending on the geographical location and heading of the directional gyroscope. Resetting is inconvenient and may result in an error if the compass happens to be under the influence of temporary accelerations when the heading is transferred from the compass to the gyroscope.

To avoid these difficulties the directional gyroscope can be reset automatically and continuously to agree with the compass.[5] Such an arrangement utilizes a remote compass transmitter and thus minimizes magnetic difficulties. Because it utilizes a directional gyroscope, its indication is stable. However, there still remains an unsolved problem in conventional directional gyroscopes, whether compass controlled or not. This is the

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represents the peak modulation (displacement in time) of a pulse from its unmodulated position. The corresponding rms displacement when the pulses are modulated fully by a single frequency test tone approaches $1/2\sqrt{2}f_c$. If S_0 denotes the corresponding rms signal voltage in the output of the multiplex, $S_0=K/2\sqrt{2}f_c$ where K is a constant factor of proportionality.

In Figure 18, N can be regarded as the voltage of an instantaneous noise sample taken at the time of triggering and ϵ as the corresponding error in timing caused by the noise sample. N can vary from sample to sample. The approximate rms value of ϵ can be taken as the rms value of N divided by SB. In this case N_0 will denote the rms noise voltage in the output of the multiplex. If the noise voltage at the input to the slicer has the characteristics of resistance noise, its rms value will equal the rms value of the instantaneous noise samples N.

Since

$$S_0=K/2\sqrt{2}f_c$$

and

$$N_0=KN/SB$$

then

$$\frac{S_0}{N_0}=\frac{S}{N}\times\frac{B}{2\sqrt{2}f_c}$$

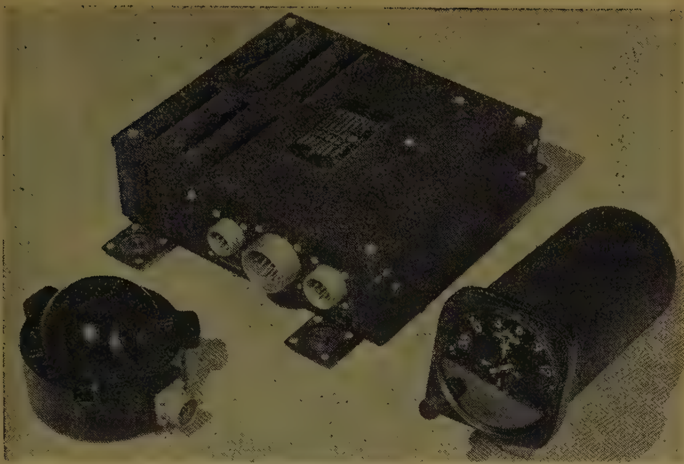


Figure 1. Components of azimuth indicating system

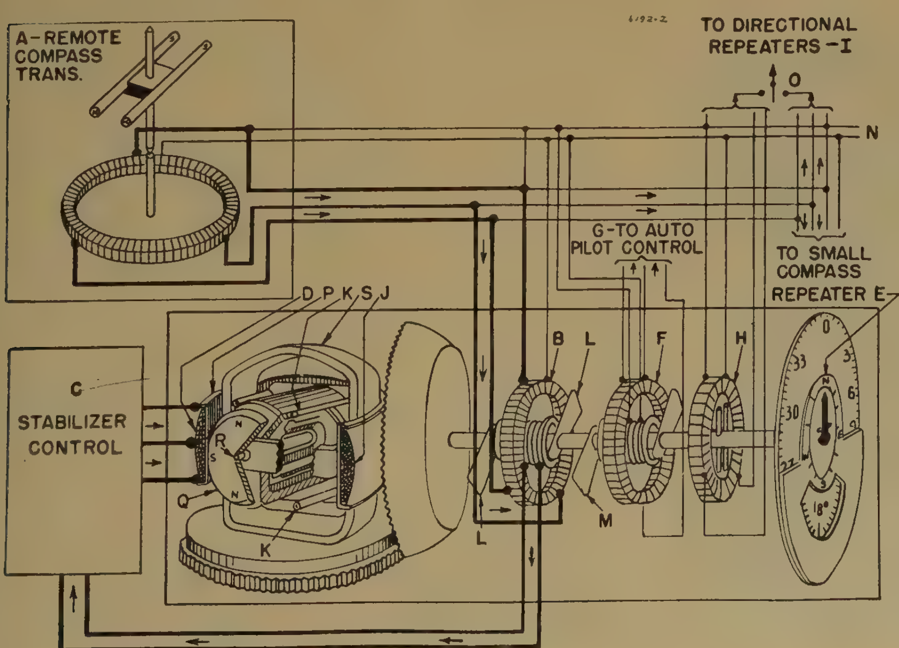
elimination of the error in indication or complete loss of indication caused by the upsetting of the gyroscope motor because of insufficient freedom of movement within its gimbal mounting. With modern aircraft violent maneuvers are possible and loops or rolls must be expected with military planes. These maneuvers will cause normal directional gyroscopes to upset.

Ideal Characteristics of a Modern Azimuth Indicating System for Aircraft

From the foregoing discussion of the limiting features in existing arrangements, it can be seen that an ideal azimuth indicator or control system for use on modern aircraft must meet certain requirements to avoid the difficulties described. Without attempting to cover all the requisites,[6] one may state that such a measurement system should be light and small, and should occupy minimum panel space. Such a system should have provision for minimizing the effects of magnetic fields or magnetic materials, and its accuracy should meet the requirements of the application. It should produce an indication which is stable. Its reading should not be excessively disturbed by normal or even violent maneuvers. It also should have some provision for minimizing turning and acceleration errors. For many applications it should have provision for transmitting the indication to a number of remote stations and it also should provide the necessary signals for azimuth stabilization or azimuth autopilot control. An additional feature considered of much importance is an emergency compass, so that if the gyroscope fails there will still remain a readable indication which is independent of local magnetic effects. A further very important feature is the characteristic of "safe failure" if anything should go wrong with the system.

Figure 2. Schematic diagram of azimuth indicating and control system

B—Selsyn detector
D—Torque motor
F—Auto pilot "pick-off"
H—Subtransmitter
J—Winding
K—Alnico bar magnet
L—Magnetic vane mounted on shaft
M—Vanes
N—115-volt 400-cycle source
O—Switch
P—Copper cylinder
Q—4-pole disk magnet
R—Motor shaft
S—Main gimbal



Important Features for an Ideal System

Three important features which help to meet these requirements are compass control, automatic gimbal adjustment, and nontumbling action. The first tends to hold the gyroscopic indication in agreement with the average compass position regardless of any natural tendency to drift. The second continuously tends to adjust the gimbals to a position at right angles to each other so that they will not drift into line and cause an upsetting condition. The third feature makes it possible to operate satisfactorily even during violent maneuvers by eliminating the undesirable tumbling action which occurs in normal limited angle directional gyroscopes.

General Description of an Improved System

The following will show how an azimuth control system, which very closely approaches the requirements of the ideal, has been obtained by the controlling or monitoring of a small nontumbling directional gyroscope by means of a remote magnetic compass. The basic devices required are shown in Figure 1 and will be discussed in detail under "Operating Principles."

In order to overcome the tendency of the gyroscope to drift in azimuth, a magnetic compass transmitter of the remote indicating type corrects the heading of the gyroscope continuously, but at a very slow rate. By this means both the natural tendency of the gyroscope to wander and the large oscillations of the compass are reduced to negligible values.

A second harmonic signal from the remote compass transmitter *A*, as shown in Figure 2, is applied to a Selsyn detector *B* located on the azimuth axis of the gyroscope. Any angular difference between the compass and gyroscope is translated

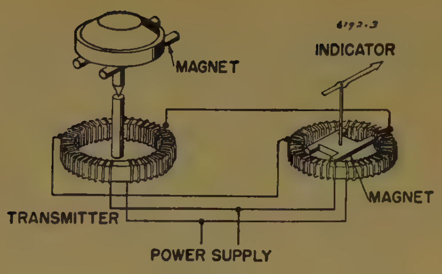


Figure 3. Basic diagram of Selsyn repeater circuits

into a voltage which is amplified in a stabilizing control box *C* and used to correct the gyroscope heading. The control is accomplished in a torque motor *D* attached to the gyroscope.

The headings of the compass transmitter and the gyroscope instrument at any instant may be compared by means of a small compass repeater *E* located in the face of the directional gyroscope. This repeater, which is connected directly to the remote compass transmitter, provides both a check on the correct functioning of the system and a very important stand-by instrument in case of a failure of the gyroscope or automatic control.

An inductive type polyphase electromagnetic "pick-off" *F* generates alternating voltages *G* capable of azimuth control of an automatic pilot or other equipment.

Because many airplanes require several simultaneous indications of direction in various parts of the airplane, a subtransmitter *H*, as shown in Figure 2, which is similar in principle to the remote indicating compass transmitter, also is provided. This is capable of operating a number of light compass repeaters *I* located as desired about the aircraft. Normally these repeaters duplicate the signal of the directional gyroscope and therefore are not subject to the usual compass errors or oscillations. Should the gyroscope fail, however, they may be connected directly to the compass transmitter by switch *O* shown in Figure 2. With this connection they provide a conventional remote compass system.

Operating Principles

A system such as the one described has many features which are responsible for its satisfactory operation. Included in a list of such features are

1. The Selsyn detector to determine the difference between compass and gyroscopic indications.

2. The stabilizer control unit that detects which direction the torque motor should run and controls that motor.

3. The torque motor for correcting the gyroscopic reading.

4. The gyroscope motor.

5. The repeater circuits which make possible the transmission of the stabilized gyroscopic indication to a large number of directional repeaters.

The operating principle of units furnishing these features will be outlined before discussing the details of the very important automatic gimbal adjustment and nontumbling action features.

Remote Compass System

The compass transmitter and repeater usually are connected in a circuit similar to that shown in Figure 3. Double frequency voltages are generated in the transmitter windings by the interaction of the d-c flux from the north-seeking compass magnets and the a-c flux from the power source in a saturatable core material. The generation of double frequency voltages by this saturating means and their use in repeater instruments already has been described adequately.[7]

The distribution of the voltages about the core is determined by the angular position of the transmitter magnets relative to the core and coils, and since the magnets are free to align themselves with the earth's magnetic field, double frequency voltages are obtained whose relative magnitudes are a function of magnetic heading of the aircraft. The usual practice is to impress these voltages upon an indicator whose magnetic construction is similar to that of the transmitter. In the indicator the interaction of the a-c flux from the power source and the double frequency flux from the compass transmitter generate a d-c field across the diameter of the indicator core. Since the angular position of this d-c field is a function of the transmitter heading, it is used to hold the indicator magnet to which the indicating pointer is attached in agreement with the transmitter heading.

Figure 5. Gyroscope motor for compass controlled directional gyroscope

Selsyn Detector

In the control of the directional gyroscope the compass transmitter is connected to the Selsyn detector. In the detector the polyphase coils and core differ from those of the conventional compass indicator or repeater by the fact that saturation of the core does not occur. Thus a double frequency flux with a direction depending on the voltage pattern of the transmitter is obtained instead

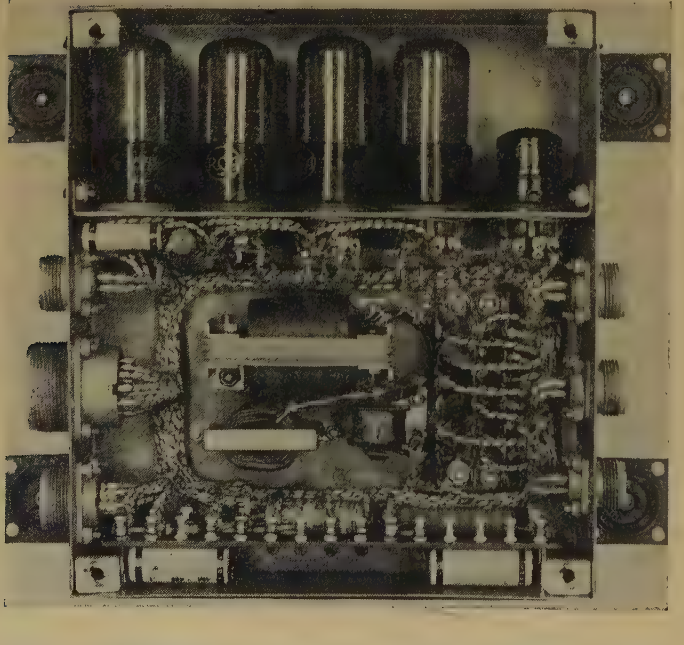


Figure 4. View of stabilizer control with cover removed

of a d-c field. Two magnetic vanes mounted on a shaft, as shown by L in Figure 2, can pick up this flux and guide it through a detector coil to detect the angular position of this double frequency flux. If the vanes are exactly in line with the a-c field, a maximum voltage will be induced in the coil, but if the vanes are turned 90 degrees from this position, zero voltage will be induced. By mounting the vanes on the gyroscope azimuth shaft the zero voltage position is used to detect when the gyroscope azimuth axis is in agreement with the compass heading. Any voltage present in the detector coil indicates an error in the gyroscope heading, the direction of which is determined by the polarity (in phase or 180 degrees out of phase from a reference voltage) of the alternating voltage.

Stabilizer Control Unit

The detector output is impressed upon an electronic control unit which amplifies and rectifies the signal to give a d-c output for operation of the torque motor to bring the gyroscope to the correct heading. A photograph of the stabilizer control unit with the cover removed is shown in Figure 4. The signal first is amplified in a conventional 2-stage amplifier circuit. It then is rectified. During rectification it is combined with a properly phased master voltage to obtain a direct current which reverses if the polarity of the input reverses. This direct current in turn is amplified and used to control the torque motor.

Torque Motor

The torque motor D, Figure 2, is operated by direct current from the stabilizer control unit circulating between the center tap and either end of a winding J. The current in the coil produces a magnetic field inside it parallel to the gyroscope main gimbal axis and in a direction determined by the direction of the correction required. Four alnico bar magnets K are placed on the sides of the motor frame with their axes parallel to the motor shaft. The magnetic field from the coil reacts with these magnets to produce a torque in such a direction that precession will take place about the azimuth axis. Precession will continue until the compass and gyroscopic indications are in agreement.

Gyroscope Motor

The gyroscope motor shown in Figure 5 is a 3-phase 2-pole hysteresis-type motor which operates synchronously at 24,000 rpm and is capable of furnishing a powerful starting torque. Adequate moment of momentum is provided by means of an external rotor shell of a tungsten alloy, and this together with the high rotating speed gives the necessary stabilizing torque. Special bearings designed to withstand high vibration accelerations give these motors improved characteristics.

Repeater Systems

In order to utilize the stabilized directional gyroscope for other purposes than as an individual indicator, additional repeater systems are employed. One of these is the subtransmitter H shown in Figure 2. In principle this transmitter is equivalent to that of the remote compass except that the transmitter magnet is attached to the gyroscope azimuth axis. By means of a circuit similar to that shown in Figure 3, the heavily damped gyroscope subtransmitter signal may be used to actuate a number of standard remote directional repeaters located conveniently about the aircraft.

A second system is the autopilot azimuth control which is initiated by a signal from a polyphase "pick-off" mounted

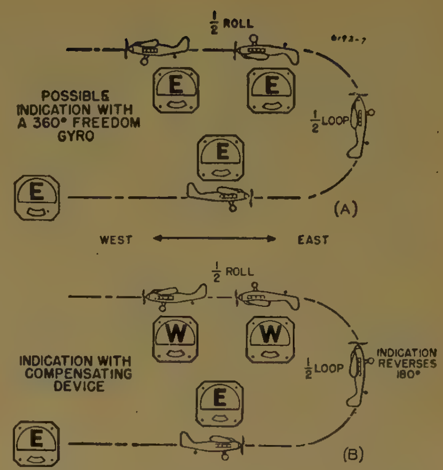


Figure 7. Indication during an Immelman turn

A. Without a compensating device
B. With a compensating device

around the gyroscope azimuth shaft as shown at F of Figure 2. The construction of this "pick-off" is similar to that of the Selsyn detector, but its operation is somewhat different. The power supply frequency is supplied to the single center coil and the vanes M distribute the flux to the outer polyphase coils in accordance with the angular heading of the gyroscope. Voltages produced in the polyphase coil are therefore a function of the gyroscope heading. These voltages may be used for accurately controlling and stabilizing various directional devices.

Automatic Gimbal Adjustment

A very important feature is used in this compass-controlled gyroscope system. An automatic method of keeping the gimbals at right angles to each other prevents the horizontal gimbal from precessing into line with the main gimbal. Such a lineup of the gimbals causes the gyroscope to upset. Figure 2 shows the construction schematically. A copper cylinder P is attached to the instrument frame in such a manner that it surrounds the gyroscope motor and its axis coincides with the main gimbal axis. In the actual design illustrated, this cylinder also serves as the mechanical support for the torque motor coil. A 4-pole disk magnet Q is attached to the end of the motor shaft R and turns with the motor rotor at 24,000 rpm. Eddy currents are induced in the copper ring by the spinning magnet and produce an electromagnetic drag acting on the motor rotor.

As it revolves the upper part of the magnet moves in one direction with respect to the cylinder, and the lower part

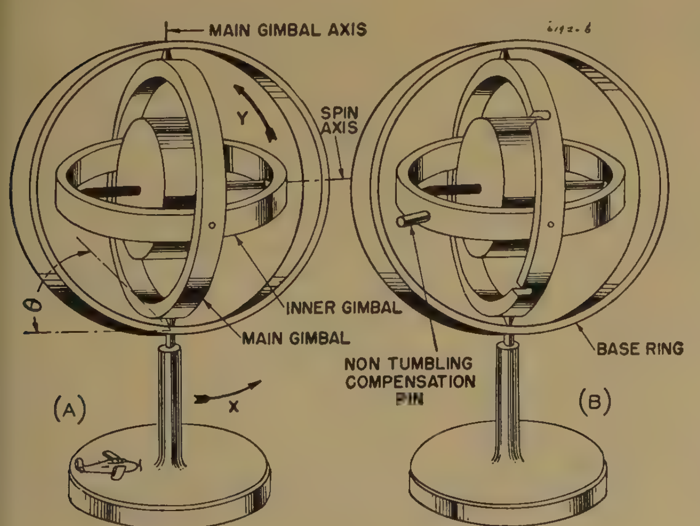


Figure 6. Typical toy gyroscope

A. Standard type
B. Special type equipped with non-tumbling compensation

moves in the other direction. The torques caused by the upper and lower parts of the magnet are, therefore, symmetrical when the spin axis is level, and no resultant drag exists about the vertical axis. When the motor axis becomes tilted with respect to the plane of the copper cylinder, the forces become unbalanced because one-half of the magnet disk is closer to the cylinder than the other. This force unbalance results in a torque about the vertical axis of the gyroscope which causes precession about the horizontal axis and prevents gyroscopic upset by keeping the spin axis at right angles to the main gimbal S.

Nontumbling Gimbal Reversal Feature

It previously has been pointed out that the system should indicate correctly after all the maneuvers that an aircraft may be expected to perform. The maneuvers include loops and rolls which cause conventional directional gyroscopes to upset because the angular movement of the spin axis in the main gimbal is limited. Universal freedom mounting can be designed to reduce the hazard of upset. A toy gyroscope such as shown in Figure 6A illustrates the general construction. Problems arise, however, which make it impractical or, at least, undesirable.

One major problem is illustrated by the fact that with 360 degrees freedom it is possible to go through certain maneuvers such as an Immelman turn where, after coming out of the maneuver, the instrument will read incorrectly by 180 degrees as shown in Figure 7A. This is further complicated by the fact that when going through a similar maneuver at a slightly different heading, this reversal will not be present.

In order to clarify this principle, consider Figure 6 which illustrates schematically a directional gyroscope. There are two conditions limiting the directional gyroscope's proper use.

1. The spin axis is fixed in direction.

2. The base is fastened to the aircraft and is upright in normal flight.

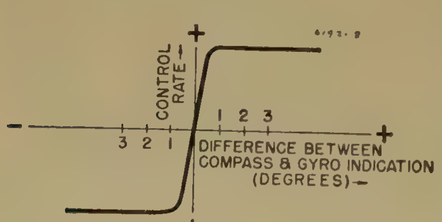


Figure 8. Typical control rate for gyroscope azimuth precession

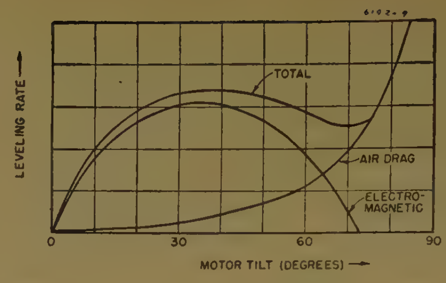


Figure 9. Typical control rate for leveling gyroscope gimbal

The angle θ, Figure 6A, between the plane of the base ring and the plane of the main gimbal is a measure of the gyroscopic indication. It is evident that with only these two limitations, the angle θ can have two values differing by 180 degrees for a particular heading of the gyroscope base ring, since besides the position illustrated, we can turn the spin axis as shown by arrow Y, continuing it through the main gimbal, and then we can bring the black end of the spin axis back to the left hand side by rotating it about the main gimbal axis. Thus for the same heading of the aircraft, we have two azimuth readings differing by 180 degrees.

The fact that both of the foregoing readings can exist is verified during an Immelman turn, which is illustrated by the following experiments. If the aircraft is considered heading in the direction shown on the gyroscope base in Figure 6A, and if the base is rotated as shown by the arrow X (keeping the spin axis in the plane of the base ring) until the base is upside down, the angle θ will not change. If the base then is returned to normal by rotating about the spin axis, θ still will be unchanged.

If the experiment is repeated in a similar manner, but if the spin axis intentionally is kept slightly off the plane of the base ring, then a phenomenon occurs which at first thought may seem strange. The main gimbal rotates 180 degrees on its own axis as the base turns upside down. This phenomenon is caused by the gyroscope's tendency to keep its spin axis pointed in one direction. When the gyroscope is mounted in gimbals with axes at right angles to each other, the main gimbal must turn in the manner described in order to allow the spin axis to hold its direction. This action is referred to as gimbal error action in less violent maneuvers.

The problem is to eliminate the possibility of having two different indications for the condition where the aircraft is upright and headed in one particular direction. A stop to prevent the spin axis passing through the main gimbal prever the possibility of two such indicatio and the arrangement is schematica shown in Figure 6B. During a loop, wh the pin hits the gimbal, it is engaged or momentarily since gyroscopic acti turns the main gimbal to free the p and therefore no undesirable tumbli action takes place.

The stop is placed so that the inn gimbal cannot quite line up with the ma gimbal, and two additional desirable r sults are attained.

1. Since the inner gimbal will not rota 360 degrees, spirals can be used to lead cu rent into the motor rather than slip rin This eliminates undesirable friction.

2. Since the axis of the inner and main gi bal never can line up exactly, the maximu undesirable effects of "gimbal lock" a eliminated. "Gimbal lock" is the conditi where the spin axis lines up with the ma gimbal axis, and in this position friction the motor bearings will tend to turn t main gimbal and thus start the direction gyroscope indicator dial spinning.

Operating Characteristics

There are certain operating characte istics of especial interest in the compas controlled directional gyroscope. Son of these may be called design characte istics since they are easily adjustable design. These would include the pr cession rate of the gyroscope azimu correction, and the leveling rate of t gyroscope gimbal automatic adjustmen

Precession Rate

The advantage of a high rate for co recting the gyroscope indication lies the faster return of the gyroscope wh large errors occur, and also in a reductio in errors caused by the drift of the gyr scope that is being compensated for, shown in the appendix. However, b cause the compass may have large erro during turns, a high rate might cau greater gyroscopic errors than a low ra Figure 8 illustrates a typical control ra curve where the amplifier is designed change from zero to the maximum cu rent very quickly, and thus give a hi correction rate for small angles of devi tion. It has an approximately consta rate for all larger angles of deviation order to reduce turn errors.

Leveling Rate

Another characteristic is the leveli rate of the automatic gimbal adjustmen With the construction shown in Figure characteristics similar to those illustrat by Figure 9 are obtained where the max mum value may be varied by the streng

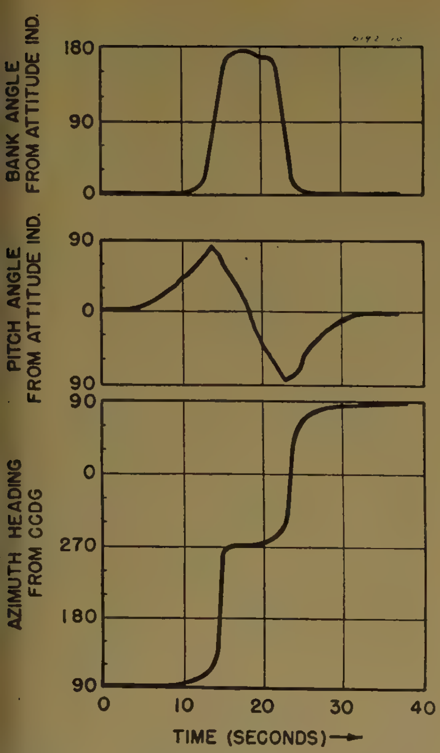


Figure 10. Data typical of actual results during a complete loop

of the alnico bar magnet. Since the dip of the motor normally is not transmitted to the dial, the shape of this curve is not particularly critical. Ordinarily there is nothing to upset the motor to a position outside the range of the copper ring, but if considered necessary, air drag on the motor rotor may be used as a leveling means for the high angles.

Flight Data During Violent Maneuvers

Very important operating characteristics are shown by actual flight data. Figure 10 shows the data taken during a loop starting on a 90-degree heading in a typical flight test.

Conclusions

This system will give satisfactory overall operation covering a wide range of voltage, frequency, temperature, and attitude conditions. For optimum performance, however, the frequency should be kept within ten per cent of its rated value, and volts per cycle within a similar range.

As fast as instrumentation and control of aircraft in azimuth have posed new problems to solve, new techniques have been developed to meet the requirements. Magnetic compasses, stabilized compasses, and compass-controlled gyroscopes have all been steps in the evolution of direction indication but have failed in the past to be adequate for the violent maneuvers which modern fast airplanes can perform. It has been shown, however, that one form of compass-controlled directional gyroscope when supplied with the new "nontumbling feature" can provide satisfactory operation for these applications. Not only can this be accomplished, but the weight and size are such that the gyroscope indicator can be mounted in a standard $3^1/_8$-inch-diameter *AN*-size instrument panel cutout, a feature which is extremely desirable.

Appendix. The Servomechanism System of an Improved Azimuth Indicator

With reference to Figure 2, the primary detector of the directional measuring system is a permanent magnet which seeks a north-south direction in the remote compass transmitter *A*. A voltage pattern is produced which is transmitted to the compass repeater *E* and, if desired, can be transmitted to the directional repeaters *I*. The main function of the voltage pattern, however, is to act as the "desired behavior" of a regulatory loop or closed casual servo system. A paper on this subject, entitled "Unified Symbolism for Regulatory Control" by G. A. Philbrick, was presented at the December 1945 meeting of the American Society of Mechanical Engineers. Figure 11 shows a graphical or symbolic diagram of the servo system.

This voltage pattern produces an a-c flux in the Selsyn detector *B* that has a direction determined by the pattern. Using the symbolism of Brown and Hall[8] this direction can be represented functionally by $\Theta_i(t)$. The gyroscope itself positions pickup vanes in the Selsyn detector, the direction of which can be represented functionally by $\Theta_o(t)$. The difference between the compass reading (corresponding to the flux direction) and the gyroscopic reading (corresponding to the vane direction) then is represented by $\Theta_i(t) - \Theta_o(t)$ and this corresponds to the error signal voltage $\xi(t)$ induced in the pickup coil.

This error signal voltage $\xi(t)$ isthe input to a stabilizer control *C*, the outputs of which are direct currents. These currents flow through the coils of a torque motor *D* to produce a torque which can be represented functionally by $T_c(t)$. If the combined operator is $C_c(p)$, then the torque $T_c(t)$ is the result of $C_c(p)$ operating on the input $\xi(t)$. This output torque $T_c(t)$ combines with disturbance torques $T_o(t)$ such as those causing drifts to cause precession of the gyroscope in a direction to reduce the error signal. This precession operator $H_o(p)$ operating on the torques determines the position of the pickup vanes.

$$H_o(p)[T_c(t)+T_o(t)] = \Theta_o(t)$$

Figure 11 easily can be recognized as a standard simple servomechanism. The operator $C_c(p)$ can be considered a constant K for small values of $\xi(t)$ because in this region the stabilizer is a proportional amplifier, and the torque from the precession coils is essentially proportional to the current. However, for large values of $\xi(t)$, the amplifier saturates, and $C_c(p)$ has the necessary value to keep $T_c(t)$ or $C_c(p)\xi(t)$ essentially constant. The differential operator $H_o(p)$ is essentially an integral operator $1/p$ since the speed of gyroscope precession, or $(d\Theta_o)/dt$, is proportional to the torque $T_c(t) + T_o(t)$.

From this discussion it is apparent that the standard simple servomechanism equations which are covered fully in another paper[8] can be used to predict errors and stability of such systems.

The three basic equations can be written as follows, omitting (t) where convenient:

$$\left.\begin{array}{l} \xi(t) = \Theta_i(t) - \Theta_o(t) \\ \text{or} \\ \xi = \Theta_i - \Theta_o \end{array}\right\} \quad (1)$$

$$\left.\begin{array}{l} T_c(t) = C_c(p)\xi(t) \\ \text{or} \\ T_c = K\xi \end{array}\right\} \quad (2)$$

$$\left.\begin{array}{l} \Theta_o(t) = H_o(p)[T_c(t)+T_o(t)] \\ \text{or} \\ \Theta_o = 1/p[T_c+T_o] \end{array}\right\} \quad (3)$$

Subtracting Θ_i from each side of equation 3 and substituting equation 2 in equation 3,

$$\Theta_i - \Theta_o = \Theta_i - 1/p[K\xi + T_o]$$

which from equation 1 gives

$$\xi = \Theta_i - 1/p\ K\xi - 1/p\ T_o$$

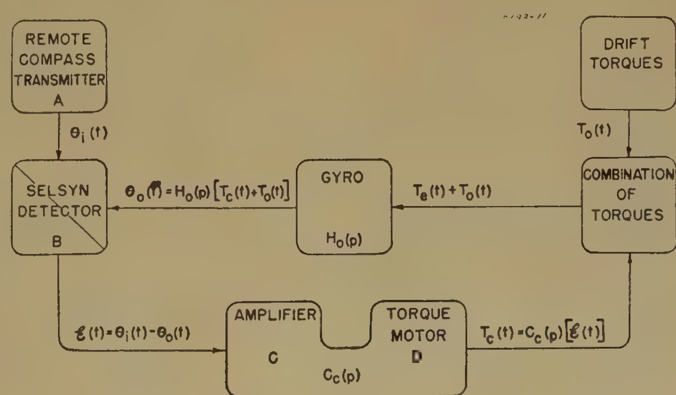


Figure 11. Graphical and symbolic diagram of compass controlled directional gyroscopic servo mechanism

or

$$\left.\begin{aligned}\xi&=\frac{\Theta_i-1/p\,T_o}{1+1/p\,K}\\&=\left[\frac{p\Theta_i-T_o}{p+K}\right]\end{aligned}\right\}\quad(4)$$

The use of this denominator as an over-all system operator has been recognized in servomechanism design for years.

Thus $(p+K)^{-1}$ summarizes the system behavior over the range that K is a constant. Since $(p+K)^{-1}$ is the Laplacian transform for ϵ^{-Kt} its characteristic is an overdamped system without oscillation and with a time constant determined by K. From equations 1, 2, and 3

$$\Theta_o(t)=\frac{C_c(p)H_o(p)\Theta_i(t)+H_o(p)T_o(t)}{1+C_c(p)H_c(p)}$$

or

$$\Theta_o=\frac{K\Theta_i+T_o}{p+K}$$

or

$$\frac{d\Theta_o}{dt}=K(\Theta_i-\Theta_o)+T_o$$

For "steady" conditions

$$\frac{d\Theta_o}{dt}=0$$

and

$$\Theta_i-\Theta_o=\xi=-T_o/K \quad (5)$$

Equation 5 leads to the conclusion that the basic error of the system can be kept small by increasing the gain represented by K and by reducing the undesired torque T_o. Equation 4 which indicates a nonoscillating system with a time constant determined by K also indicates that K may be large without producing undesirable characteristics.

If K is to be large over the range that it is a constant, and if the precession rate is to be limited beyond this range in order to reduce errors that result from the gyroscope following the compass during the period that turn errors are present in the compass, a characteristic curve as shown in Figure 8 results.

With this characteristic, errors caused by normal undesirable torques are negligible and normal turn errors are small. If existing turn errors are objectionable, the compass control can be disconnected during turns.

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Some Aspects of Polyphase Motor Design—The Design and Properties of the Magnetic Circuit

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IN a preceding paper,[1] an attempt was made to present a design point of view, indicating that the criterion for a design method is its direct ability to provide the construction data whereby desired machine specifications are met. The idea was illustrated by a procedure for polyphase motor design, stripped of refinements, and employing a magnetic circuit of fixed contour. Such a condition arises through the use of standard laminations in which designs are made. The questions naturally arise as to how these laminations are proportioned and what affect the proportions have on performance. It is with such matters that the present paper is concerned, directed chiefly at motors with semiclosed slots of 50 horsepower or less.

An investigation of the various proportions and dimensions which can be used in the magnetic circuit of an electric motor leads to some interesting relationships. Unfortunately, these readily can become of academic rather than practical interest unless several underlying facts of industrial and commercial importance are kept in mind.

Faced with the requirements of designing a suitable stator and rotor lamination for one definite horsepower and speed rating for a polyphase induction motor, it can be imagined readily that the relation of slot area to tooth area, and rotor diameter to stator diameter (and several other proportions) could be set up in useful ratios so as to obtain minimum copper, or minimum material cost, maximum efficiency, or any one of several other desideratum. This is a process of designing a lamination for a particular rating. In contrast to this is the process of designing a lamination for a given frame. The distinction is important enough to bear further explanation.

It has been pointed out that while a motor manufacturer sells horsepower, in reality he sets up to build frames. In some instances, a special frame or a special design may be made for one application, bearing one rating only, and planned for very large production. In such a case, the design of a "best" lamination for the case at hand may be justified.

In the majority of cases, a lamination is to be provided for a given frame and made as widely useful as possible for all ratings which expect to be sold in that frame size. Suppose, for instance, NEMA* frame 284 is considered. Standardized ratings in Table I indicate the horsepower and speeds which all are expected from this

Table I

Horsepower	RPM	Horsepower	RPM
General purpose, open type, 60 cycles			
10	3,600	1 1/2	600
7 1/2	1,800	1	514
5	1,200	3/4	514
3	900	1/2	450
2	720		
Totally enclosed			
3	1,800	2	900
General purpose, open type, 25 cycles			
5	1,500	2	500
3	750	1 1/2	500

* National Electrical Manufacturers Association.

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"container" of more or less fixed dimensions.

From a commercial point of view, all ratings shown should be obtained from exactly the same lamination, stacked to the same axial length. Such a practice would mean that any order for such ratings could be filled from a stock of built-up unwound bodies and fields, modifying only the stator windings and the end ring sections of the squirrel cage. This ideal state cannot be realized but it can be approximated, and the attempt has a great effect upon lamination design and the functions of the designer.

Considering the foregoing facts, an attempt to design a "best" lamination can be of interest only in the exceptional case. If the lamination is to be widely useful, the proportions determined as best for one rating readily may prove undesirable for another. The result is a series of compromises, based on the minimum number of laminations for any given frame, which will yield all of the standardized (and some special) ratings as required, without serious interference with the accepted standards of performance.

Dealing with lamination contours for such purposes, it is reasonable to take a pragmatic viewpoint. Accordingly, we will investigate first the proportions of a series of laminations which, through refinements of practice, have been known to yield useful results.

Method of Fixing Stator Proportions

A total of 108 stator and rotor laminations were used to determine lamination proportions. These varied in rotor diameter from 3 to 15 inches. Their proportions had been determined initially by investigations or as manifestations of the individual tastes of a number of designers over a period of years. As a first step in determining proportions, a series of points were plotted using diameter to the bottom of the stator slots versus rotor diameter. With surprisingly little variation these points fell on a straight line which can be expressed by the equation (see Figure 1)

$$D_1 = 1.175D + 0.647 \quad (1)$$

To determine the outside diameter of the stator, relative flux densities of teeth and stator yoke will be used. Maximum tooth densities and average yoke densities will be determined.

Maximum stator tooth density is

$$B_{max} = \frac{1.57 \times \text{poles} \times \text{flux per pole}}{\text{number of slots} \times t_w \times L} \quad (2)$$

The tooth width (t_w) is assumed to be the narrow straight portion shown in Figure 1.

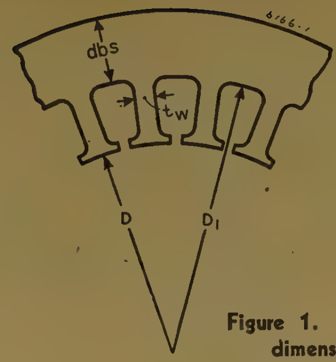

Figure 1. Stator dimensions

Average density in the yoke is

$$B_{ave} = \frac{0.5 \times \text{flux per pole}}{\text{radial yoke length} \times L} \quad (3)$$

The radial yoke length or depth below slot will be abbreviated *dbs*.

As a working basis, it will be assumed that a reasonable magnetic circuit results from proportions giving a maximum stator tooth density of 120 per cent of the average yoke density. This ratio can vary somewhat with few bad effects. It is obvious that if tooth densities were kept constant on say both 4- and 8-pole designs, the yoke densities would reduce to half in the latter case. As will be shown in dealing with compromise laminations, over-all densities may well be adjusted. The effect of other ratios is investigated briefly in the appendix. From the 120-per-cent ratio,

$$\frac{1.57\phi P}{S_1 t_w L} = 1.20 \frac{0.5\phi}{L(dbs)} \quad (4)$$

Solving for depth below slot,

$$dbs = 0.382 \frac{S_1 t_w}{P} \quad (5)$$

Twice the depth below slot, plus D_1 represents the outside diameter of the stator laminations. It would be convenient to obtain this in terms of rotor diameter, and to do so the following method was employed.

The portion of the stator lamination lying just within the circle swept out by D_1 obviously fulfills the double purpose of supplying both the magnetic and the electric circuits. The division of this ring between slots and teeth affects the performance and results in an iron or a copper design. The effect is illustrated briefly in the appendix, but at this point it is sufficient to point out that a motor requiring low current densities (as a totally enclosed design with a cooling problem) might require a greater portion of this area for copper. Conversely, a design requiring high maximum torque, and intended for intermittent service, can be shown to require an increased tooth width. Our interest, however, is to determine a working value which is useful for general purpose designs, and hence the average for a group of existing laminations were investigated. When ($S_1 \times t_w$) versus rotor diameters were plotted for the 108 laminations, the results spread over a band which gave a clearly defined straight line. The equation is

$$S_1 t_w = 1.35D \quad (6)$$

Substituting in equation 5,

$$dbs = 0.515 \frac{D}{P} \quad (7)$$

Then

$$OD = 1.175D + 1.03 \frac{D}{P} + 0.647 \quad (8)$$

Solving for D,

$$D = \frac{OD - 0.647}{1.175 + \frac{1.03}{P}} \quad (9)$$

This equation, along with equation 1, is plotted in Figure 2 against rotor diameter. Based entirely upon average relationships observed from a number of laminations used in practice, and a few fundamentals, it fixes the main proportions of a stator lamination. Its validity as a working method will be checked later.

Compromise Dimensions

To illustrate the use of these curves or formulas, assume that laminations were to be provided for a frame which could accommodate a lamination diameter of ten inches. By these rules, Table II can be made.

Table II indicates that for each number of poles a different lamination would be necessary, and hence this practice represents the very requirement which is least desirable commercially. The point is, however, that by an examination of the dimensions, progressing in a logical manner, an intelligent compromise of dimensions can be made. Any such compromise

Table II

Poles	D, In.	D_1, In.	dbs, In.
2	5.53	7.14	1.42
4	6.53	8.32	0.84
6	6.94	8.80	0.595
8	7.16	9.06	0.46
10	7.32	9.25	0.377
12	7.42	9.36	0.317
14	7.49	9.45	0.276
16	7.55	9.52	0.242
18	7.59	9.56	0.217
20	7.62	9.60	0.196

involves a grouping of numbers of poles, and a consideration of mechanical strength of the yoke, in cases of many poles.

A 2-pole lamination usually represents an isolated design, not used for any other speed. This distinction is obvious, as the flux density in the yoke is only half of that obtained for a 4-pole machine, if the two designs are attempted in the same lamination. This change is always proportionately less as the poles increase in number. Table II illustrates this point.

No absolute rules can be laid down as to the grouping to be followed, partially because the relative production in various lines represents an individual problem. Thus compromise dimensions might be set up between the 4- and 6-pole designs, planning all production of a greater number of poles in one lamination. However, it might prove feasible to plan 4-, 6-, and 8-pole production in a compromise lamination, using a different design from 10-pole up. If 36 stator slots are used, fractional slot winding obviously would be required on the 8-pole designs. Grouping the 8-pole requirements with the lamination designed for the greatest number of poles, and using 72 slots, results in comparatively few fractional slot windings from 8 to 20 poles. Three-phase cases are emphasized here.

Granted that the grouping is an individual problem, it is of interest to observe the general effects on performance which would result in using a lamination designed by these rules for 4-pole use, when applied to 6-pole motors and vice-versa. Such results are tabulated in the appendix and are contrasted with other designs made in compromise laminations.

Stator Tooth Shape

Shape of the tooth tip and contour of the slot bottom are necessary for the completion of the stator lamination design.

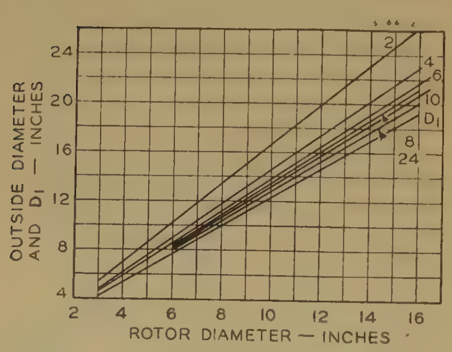

Figure 2. Values of outside diameter and D_1

The former varies with the number of poles when plotted as a function of rotor diameter

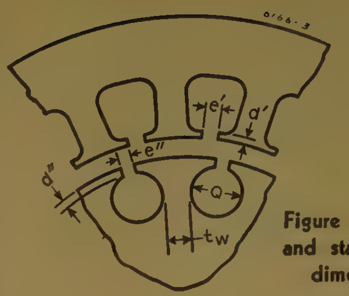

Figure 3. Rotor and stator tooth dimensions

sign. Relative values of d' and e' are important to the performance (see Figure 3). Too small a value of e' may make insertion of the windings difficult or impossible; too large a value increases the tooth flux fringing, the effective gap length, and the surface losses. An increased value of d' shows up chiefly in the increased slot leakage reactance.

A suggested working basis is to calculate

$$e' = 0.0143D + 0.0643 \tag{10}$$

In designs with an unusually great number of slots for comparatively small diameters, the value of e' given by this formula may prove quite large. Any modification should be made towards reducing this value.

The depth d' should be chosen so that

$$d'/e' = 0.3 \text{ (approximately)} \tag{11}$$

The bottom of the slot can be made semicircular, tangent to the circle swept out by D_1; or it can be made flat, with rounded corners. No recommendations can be made on this choice, aside from the comment that a die maker usually prefers the former practice.

Area of Stator Slots

The toroid between D and D_1 represents an area from which total tooth area can be subtracted to give gross slot area. Such a calculation gives the following expression for gross area of all slots:

$$\text{Area (sq in)} = 0.18D^2 + 0.758D + 0.33 \tag{12}$$

Note that this is independent of the number of slots. Actually, because of rounding off corners, the net useful area per slot for laminations designed by these rules is approximately

$$\frac{0.9 \text{ gross area}}{S_1} = \text{actual slot area} \tag{13}$$

Gross area is plotted in Figure 4.

Rotor Lamination Design

Items to be considered in the design of squirrel cage rotor laminations are the number and size of slots, and the dimensions of the slot openings. Common practice makes use of the following two forms of construction:

1. The round slot used with copper bars.
2. The deeper slot used with die cast cages.

The general problem will be considered in more detail for the first case.

It already has been pointed out that the stator slot opening can be obtained through the empirical relationship of equation 10. A useful value for the rotor opening is 70 per cent of e' or

$$e'' = 0.01D + 0.045 \tag{14}$$

Modification usually should occur on the low, rather than the high side.

The slot constant K_s'' expresses the effective linkages per ampere per inch of length for a given slot shape. As a working basis, it will be assumed that this should always be 1.3. Now, since

$$K_s'' = 0.623 + d''/e'' \text{ (for round bars)} \tag{15}$$

it follows that

$$d''/e'' = \frac{d''}{0.01D + 0.045} \text{ or } 0.677$$

Hence

$$d'' = 0.00677D + 0.0304 \tag{16}$$

Both d'' and e'' are now fixed by these equations

The diameter of the slot remains to be determined. A working basis for deciding on slot diameter, and hence tooth width, involves a consideration of flux densities in the teeth. For field laminations the tooth width was fixed thus:

$$t_w = \frac{1.35D}{S_1}$$

In the same manner, if equal flux densities in the stator and rotor teeth are desired,

$$t_w'' = \frac{1.35D}{S_2} \tag{17}$$

where S_2 = the number of rotor teeth or slots.

When round slots are used, the assumption that t_w'' indicates the tooth width at the narrow neck usually permits higher densities in the rotor than in the stator (see Figure 3). A figure of 118 per cent is useful and hence

$$t_w'' = \frac{1.35D}{1.18S_2} \tag{18}$$

For cast rotors in which the tooth is commonly longer, equal flux densities in stator and rotor teeth are more desirable.

The magnetizing ampere turns required for rotor teeth otherwise may become excessive.

While the foregoing relationship enables the designer to determine the tooth width, a more direct calculation for fixing the diameter of the slot (and hence the bars) is very useful. This problem will be instigated further.

Our first concern is with an expression for the slot or bar circle diameter. A distinction must be made concerning diameter D as to whether it is used as the true rotor diameter or the stator bore. If considered as the stator bore, the air gaps immediately enter the calculations. While practices as to suitable gap lengths vary considerably, a reasonable basis is

$$\Delta = 0.0016\times D + 0.0072 + 0.001L \qquad (19)$$

If modified, this usually would be increased for higher speed motors.

This gives the radial length of air gap in inches, as a function of both diameter and stack length. For our purposes it will be sufficiently accurate to assume a gap resulting from an average stack of three inches, leaving gap length as a function of diameter only.

To determine the slot diameter, we can make use of the following relationship:

$$\text{Diameter across slot centers} = D - 2\Delta - 2d'' - Q \qquad (20)$$

where Q = diameter of slot in inches. (This equation is in error to the extent that d'' is not measured on the diameter D, but is offset by the amount of one-half e''.)

Expanding equation 20, by equations 19 and 16 we obtain

$$\text{Diameter across slot centers} = 0.9833D - 0.0752 - Q$$

From which it follows that the diameter of the slot is

$$\frac{1.95D - 0.236}{S_2 + \pi} = Q \qquad (21)$$

It is of interest to note how the total area of the rotor slots varies as a function of their number. These areas are shown in Figure 5.

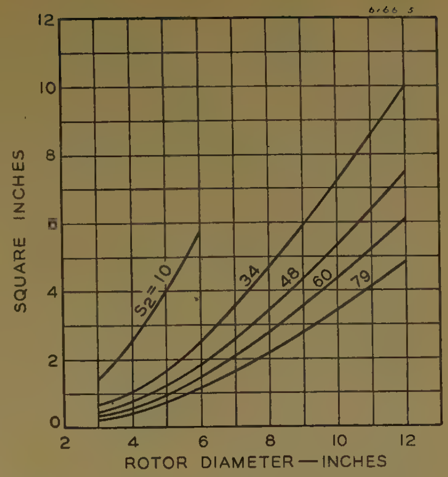

Figure 5. Total area of all rotor slots for various numbers of slots

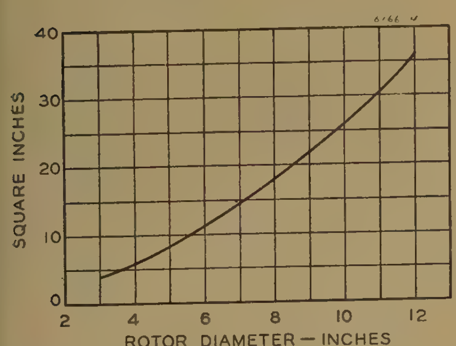

Figure 4. Total area of all stator slots as a function of rotor diameter

Recapitulation

The foregoing rules are all that are required to determine the dimensions for stator and rotor laminations, once the outside diameter and the number of poles are known. Yet, it must be kept in mind that perhaps the greatest use of such formulas occurs through the design of general purpose or average laminations which are intended for use for a variety of designs, involving various numbers of poles. The assumption is that if one wishes to design a lamination for 4-, 6-, and 8-pole use the procedure would involve the preliminary design for 4- and 8-pole, and thus make an intelligent compromise between the known limits.

No data on recommended numbers of slots are given here as that involves other factors not relevant to this paper.

Lamination Capacity

It cannot be emphasized too strongly that once a set of laminations has been proportioned the maximum torques available from all motors built with these laminations have been fixed. Aside from the slight variation possible by a change in pitch, or the working of the magnetic circuit at various densities, a given stack of laminations results in a fixed value of maximum torque.

These facts give rise to the interesting possibilities of determining the inherent maximum torque of a given lamination when designed by the method evolved here and when designed according to any or no method.

It is possible to set up expressions for all motor constants in terms of the dimensions resulting from the rules given here. Then, by selecting a flux density for the teeth, the maximum torque for an entire line of laminations can be determined. Such calculations, while not shown here, can be used as the basis for empirical equations for maximum torque as functions of stack length and rotor diameter. By such a method, 4-pole 60-cycle designs of full pitch windings, and with a maximum stator tooth density of 96,000 lines per square inch, display maximum torques of

$$T_m = KL^{1.4}D^{1.87} \qquad (22)$$

where stator and rotor slots are, respectively, 24 and 34, K equals 0.21; 36 and 48, K equals 0.31; and 48 and 60, K equals 0.40.

Similarly, for 2-pole motors designed in these laminations according to the foregoing standards,

$$T_m = KL^{1.59}D^{1.63} \qquad (23)$$

where stator and rotor slots are, respectively, 10 and 12, K equals 0.10; 24 and 34, K equals 0.34.

These constants will vary as the square of the ratios of the densities, for any other values chosen.

It is of interest to compare the result of these empirical expressions with the actual design carried through in some detail in Appendix I. A maximum torque of 52.9 pound-feet was obtained therein. By equation 22

$$T_m = 0.31\times 3^{1.4}\times 6.875^{1.87}$$
$$= 53.5 \text{ pound-feet}$$

Maximum Torque of Any Lamination

Calculation of the maximum torque of any polyphase induction motor involves the determination of the stator winding resistance, the leakage reactance, and the leakage factor. Basically, these all are calculated from certain physical dimensions of a lamination stack and the windings which can be provided in the laminations.

Such expanded expressions obviously can be substituted in the maximum torque equation to obtain an equation in terms of dimensions. As this substitution is straightforward, it requires no derivation here. The final equation is

$$T_m = 0.029\left(\frac{1-P^2K_a}{K_d}\right)\frac{B_k^2L^2}{K_aP}\times \frac{1}{\frac{K_g}{f}\left(\frac{K_h}{P}+L+K_c\right)+K_aL+\frac{K_m}{P^2}} \qquad (24)$$

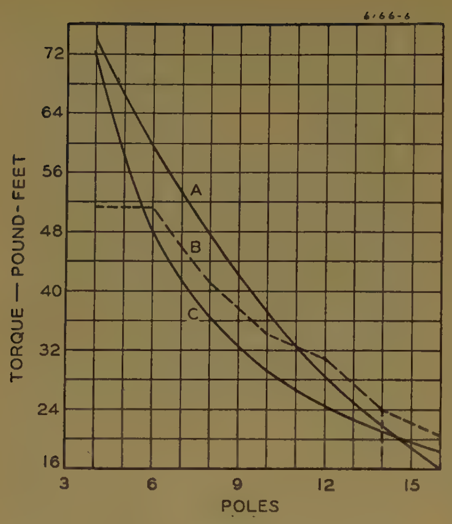

Figure 6

A—Inherent maximum torque of a lamination with constant flux density in the teeth

B—Maximum torque of standardized NEMA ratings (225 per cent ratio)

C—Inherent volt-amperes of a lamination with constant current densities. Scale not shown, as only the relative slopes and positions of the curves are significant

where

$$K_a = 3.60(0.625p + 0.375)\left(\frac{K_s'}{S_1} + \frac{K_s''}{S_2}\right) + \frac{7.5DK_z}{S_1 S_2 \Delta}(0.8p + 0.2)$$

$$K_z = \left(\frac{t' + t''}{\lambda' + \lambda''} - 0.5\right)^2$$

$$K_c = 0.2D$$

$$K_d = 0.845\frac{DK_w^2}{\Delta K_1}$$

$$K_1 = \frac{\lambda' \times \lambda''}{t' \times t''}$$

$$K_e = \frac{328}{(K_w t_w S_1)^2}$$

$$K_g = \frac{0.153 \times 10^8}{AkS_1}$$

$K_h = 4.4p(D + \text{stator tooth length})$
$K_m = 1.57D(1.25p - 0.25)$

and

P = number of poles
p = pitch as a decimal fraction
B_k = maximum flux density in the stator tooth in kilolines per square inch
f = frequency in cycles per second
$K_s' + K_s''$ = primary and secondary slot constants, respectively
$S_1 + S_2$ = number of primary and secondary slots, respectively
Δ = the radial air gap length in inches
$t' + t''$ = the effective tooth face in inches for primary and secondary, respectively. These are wider than the actual circumferential tooth widths by the effect of fringing
$\lambda' + \lambda''$ = the tooth pitches of primary and secondary, respectively
K_1 = the air gap constant. Greater than unity to allow for fringing effect
t_w' = the effective stator tooth width in inches
A = slot area in circular mils
k = the space factor for the slot (0.25 to 0.30)

Fortunately, these constants are properties of a given lamination and need not be recalculated for various designs. Specifically, they are unaffected by changes in stack length, frequency, flux density, or voltage. If the change in winding factor with the number of poles is neglected in making rough calculations for estimates, the constants are unaffected. To be more exact, K_a, K_d, K_e, K_h, and K_m must be adjusted for each change in pitch and winding factor.

A detailed example of this calculation is given in Appendix IV. However, its significance can be shown by assuming that one were interested in investigating design possibilities of everything from 4- to 16-pole motors in exactly the same lamination and stack. The inherent maximum torque of the lamination used in the appendix is shown plotted in Figure 6. The relation of such torques to the torque ratings required of the motors built in a corresponding NEMA frame also are indicated by these curves. Minimum ratios of maximum torque to full load torque are fixed at 200 per cent by NEMA, but the plotted values are at 225 per cent.

Volt-Ampere Rating of a Lamination

The volt-ampere capacity of a lamination is a fixed quantity only if allowable flux and current densities are chosen. The relationship can be derived as follows:

$$\text{Circular mils per conductor} = \frac{AkS_1}{nC} \tag{25}$$

For a given maximum flux density in the teeth of B_k kilolines,

$$C = \frac{1}{L}\left(\frac{EP \times 70.9 \times 10^3}{B_k f S_1 t_w K_w}\right) \tag{26}$$

All terms have been defined previously.

If the allowable current density in circular mils per ampere is D_c, it follows that

$$\text{Volt-amperes (all phases)} = \frac{AkS_1^2 LB_k t_w f K_w}{D_c P \times 70.9 \times 10^3} \tag{27}$$

The foregoing formula holds for any lamination, but if the given rules are followed, the following simplifications result for laminations designed accordingly. Since

$$t_w = \frac{1.35D}{S_1}$$

$$\text{Volt-amperes} = DL\frac{AkS_1B_k fK_w}{D_c P} \times 1.9 \times 10^{-5} \tag{28}$$

Assuming an average space factor for the slot of 0.28, useful slot area can be expressed by the approximate relationship

$$AkS_1 = 0.28 \times 0.9 \times D^{1.41} \times 1{,}273{,}000$$

Then

$$\text{Volt-amperes} = D^{2.41}L\frac{B_k}{D_c}\text{rpm} \times K_w \times 0.051 \tag{29}$$

Comparing this equation with the empirical expression for maximum torque, it becomes clear that as the diameter of the rotor (or correspondingly, the outside diameter of the laminations) increases,

Table III. Four-Pole Design

Lamination Outside Diameter 10.5 Inches, Slots 36/48, Stack 3 Inches
Full Load Torque: 7 1/2-Horsepower 4-Pole Motor, 22.9 Pound-Feet

	1	2	3
Nominal rotor, D	6.875 in.	7.75 in.	6.5 in.
Depth below slot, *dbs*	0.885 in.	0.84 in.	1.0 in.
Tool width, t_w	0.257 in.	0.30 in.	0.225 in.
Slot constant, K_s'	1.295 in.	0.857 in.	1.3 in.
Slot area (sq in.)	0.355 in.	0.184 in.	0.400 in.
Wire weight (lbs)	15.6	8.1	17.6
Stator tooth density	96,000	96,000	96,000
Stator yoke density	80,000	98,000	62,000
Stator winding resistance	0.452 in.	0.64 in.	0.525 in.
Rotor slot constant, K_s''	1.3	1.3	1.3
Maximum torque	52.9	57.0	42.6
No-load values			
Amperes	6.3	8.0	5.04
Copper loss	53.6	123.0	40.0
Iron loss	150.8	178.0	128 7

Notes: Column 1—"Average" method.
Column 2—Large D, t_w.
Column 3—Small D, t_w.

the volt-ampere capacity of the laminations increases more rapidly than the maximum torque. The inverse of this has been noted by any designer of small polyphase motor parts, which are generally capable of yielding comparatively high maximum torque values. Attempting to give such parts their corresponding horsepower ratings can be done only at a sacrifice of circular mils per ampere.

The motors used in Appendix IV are calculated for inherent volt-ampere capacity and plotted in Figure 6. An allowance of 425 circular mils per ampere was used for the stator winding.

Appendix I

To illustrate the application of the formulas for the average lamination, assume that a lamination is to be designed for an outside diameter of 10.5 inches. As a working basis, the maximum stator tooth density of 120 per cent of average yoke density will be used.

From equation 7

$D = 6.875$ inches (4 poles)

Selecting 36 stator slots, we obtain the tooth width from equation 5:

$t_w' = 0.257$ inch
$dbs = 0.885$ inch

From equation 10, we obtain the slot opening:

$e' = 0.162$ inch

For $d/e = 0.3$,

$e = 0.049$ inch

"Rounding off" the corners of the slots yields a slot area of 0.355 square inch. This completes the stator lamination design, determined not as a fair lamination for say both 4- and 6-pole designs, but based on 4-pole calculations only.

For the rotor, using 48 slots and a gap of 0.025 inch (which is subtracted from the nominal diameter), the diameter of the rotor slot is 0.255 inch.

For semiopen slots, e'' equals seven-tenths of e', or 0.11 inch; and d'', when the rotor is ground, will be such that the slot constant is 1.3. The tooth width at the narrow portion is now 0.164 inch.

These laminations will be used for a design of a $7^1/_2$-horsepower motor, using the principles derived in the first paper on this subject.[1] Selecting a maximum to full load torque ratio of 225 per cent, and a suitable stator tooth density as 96,000 lines per square inch, the required stacking will be 3.0 inches.

For a full pitch winding, the complete performance is given in Table III, column 1.

The solution of the foregoing problem represents purely a routine task, of little significance beyond the indication that the lamination design method results in reasonable values. It leaves unanswered the following questions:

1. What would happen had a larger or a smaller rotor diameter been chosen?

2. What would be the effect of upsetting the ratio set up between tooth and slot areas in the stator?

To investigate, assume that a rotor diameter of 7.75 inches were chosen, along with a tooth width of 0.30 inch. The flux densities in stator core and teeth are now practically equal and are kept at 96,000 lines per square inch, as before. The lamination design, in general, is what would follow if 1.2 were changed to 1.0 in equation 4. The completed performance is tabulated in column 2 of Table III and bears out the conclusion that such a procedure results in a "flux motor" with high maximum torque, but overworked copper. Such a lamination design may represent a cooling problem, but might be useful for intermittent duty service with high torque demand.

In the other direction, suppose a lamination had been designed for a rotor diameter of 6.5 inches with a tooth width of 0.225 inch. The attempt to limit the flux density in the stator teeth to 96,000 lines now results in a weak motor, with maximum torque below the 200 per cent ratio if a $7^1/_2$ horsepower rating is expected. The increased slot area naturally results in less copper loss. Performance items are shown in column 3 of Table III.

Table IV. Six-Pole Design

Laminations Outside Diameter 10.5 Inches, Slots 36/48, Stack 3 Inches
Full Load Torque: 5-Horsepower 6-Pole Motor, 22.9 Pound-Feet

	1	2	3
Nominal rotor, D	7.315 in.	6.875 in.	7.315 in.
Depth below slot, dbs	0.628 in.	0.885 in.	0.628 in.
Tooth width, t_w	0.274 in.	0.257 in.	0.274 in.
Slot constant, K_s'	1.28	1.295	1.28
Slot area (sq in.)	0.386	0.355	0.386
Wire weight (lbs)	14.3	12.6	14.3
Stator tooth density	96,000	96,000	98,000
Stator yoke density	80,000	53,300	81,700
Stator winding resistance	0.659	0.814	0.630
Rotor slot constant	1.3	1.3	1.3
Maximum torque	50.2	42.9	52.6
No load values:			
Amperes	6.42	5.83	6.68
Copper loss	72.0	85.2	84.0
Iron loss	130.0	111.8	134.0

Column 1—"Average" method for 6-pole lamination.
Column 2—"Average" 4-pole lamination with 6-pole design.
Column 3—6-pole lamination with redesign of windings for higher value of maximum torque.

Appendix II

Examples given previously were based on the use of a lamination intended primarily for 4-pole operation. Using the same principles, but designing the lamination for 6-pole operation only, results in a rotor diameter of 7.315 inches with a tooth width of 0.274 inch. The resulting performance items are shown in Table IV, column 1. The 6-pole rating in this same stack represents a 5-horsepower motor. It will be noted that this motor is somewhat weaker in torque ratios than the one designed in Appendix I. A slight increase in densities brings about an equivalent value of maximum torque, with performance as shown in column 3.

While the data are not shown, a modification of this "average" lamination to yield equal densities in stator yoke and teeth, or to increase the tooth densities beyond the 120 per cent ratio setup, results in exactly the same trends already indicated in Appendix I.

In attempting to judge the validity of these lamination designs, one of the most effective means is to determine the effect of not following the rules. Thus, while average laminations have been designed for 4- and 6-pole use, it would be of interest to show how satisfactory a 6-pole design could be made in the lamination intended for 4-pole service.

The lamination of Table III, column 1 yields a 6-pole design as shown in Table IV, column 2. A flux density of 96,000 lines in the stator teeth again was chosen as a basis for calculation. As compared to the "average" lamination for 6-pole use, a sharp reduction in maximum torque results, along with an increase in copper loss.

This comparison bears out the fact that as an increased number of poles is used with a given lamination, the available torque always is reduced so long as tooth densities are maintained constant. This effect is exaggerated here, because the laminations had been deliberately designed for 4- or 6-pole duty respectively. In practice, the contrast would not have been so marked if the lamination dimensions had been compromised between 4- and 6-pole requirements, or if constant tooth densities were not maintained. Since an increase in the number of poles always results in reduced stator yoke densities (on a given lamination), an increase in tooth density then is not unreasonable. This may result in no great increase in core loss nor magnetizing current, and this practice does reduce the slope of the maximum torque versus number of poles curve for a given lamination.

Appendix III

It has been indicated in Appendix I that a 4-pole lamination for the outside diameter required should have a rotor diameter of 6.875 inches, and that the resulting performance was reasonable. Similarly in

Table V. Four- and 6-Pole Designs

Laminations Outside Diameter 10.5 Inches, Slots 36/48, Stack 3 Inches

	4-Pole		6-Pole	
	1	2	3	4
Full load torque	22.9 lb-ft	—	22.9 lb-ft	
Inside stator bore	7.00 in.	—	7.00 in.	
Depth below slot, *dbs*	0.826 in.	—	0.826 in.	
Tooth width	0.265 in.	—	0.265 in.	
Slot constant, K_s'	1.31	—	1.31	
Slot area	0.34	—	0.34	
Wire weight (lbs)	15.1	15.1	12.2	12.2
Stator tooth density	96,000	97,500	96,000	104,000
Stator yoke density	88,000	89,500	59,000	64,000
Stator winding resistance	0.374	0.385	0.708	0.602
Rotor slot constant, K_s''	1.3	1.3	1.3	1.3
Maximum torque	51.2	53.0	43.8	51.0
No load values:				
Amperes	6.6	6.8	6.1	7.3
Copper loss	51.0	55.0	83.0	101.0
Iron loss	155.0	161.0	116.0	129.0

Appendix II, a 6-pole lamination has been shown with a nominal diameter of 7.315 inches. Performance was acceptable, but when the 4-pole lamination was used on a 6-pole design, results were less satisfactory.

To illustrate that these rules are of aid in achieving a compromise design, let us assume that one lamination were to be used for both, and that its dimensions are selected as being between the two extremes.

Between nominal diameters of 6.875 and 7.315 inches, we will select 7 inches for the stator bore.

Between stator tooth widths of 0.257 and 0.274 inch, select 0.265 inch which follows from equation 6.

By equation 1, $D_1=8.847$ inches and $dbs=0.826$ inch.

The 4- and 6-pole designs made in this compromise lamination are tabulated in Table V. The significance of the statement that compromise laminations require adjustments in flux densities is apparent at once in examining columns 1 and 3 of this table. While the 4-pole motor suffers slightly in maximum torque, the 6-pole design is below the 200 per cent minimum, if the 96,000 density is used. Columns 2 and 4 show the results of adjusting the designs for the values of maximum torque obtained previously. The resulting performances do not compare too unfavorably with those obtained from the laminations made for individual designs

Appendix IV

To illustrate the application of the inherent maximum torque of a lamination, we will consider a sample lamination designed without reference to the rules shown here, and displaying the following properties:

Outside diameter = 11.75 inches
Inside diameter = 7.50 inches
Slots = 48
Useful slot area = 123,000 cubic centimeters
Slot constant = 1.94
Tooth width = 0.23 inch
Tooth length = 1.125 inches
Rotor outside diameter = 7.445 inches
Air gap = 0.0275 inch
Slots = 60
Slot constant = 1.3
Gap constant = 1.27
B_k (assumed) = 85
Stack length, $L=3.00$ inches

Assume that this set of laminations was used for designs from 4 to 16 poles inclusive, using a fractional slot winding in the necessary cases. The initial calculations will be made with a pitch of 0.75 and a winding factor of 0.88 used as constant throughout.

Then

$K_a=0.271$
$K_c=1.50$
$K_g=2.60$
$K_m=8.10$
$K_d=141.0$
$K_e=3.45$
$K_h=28.5$

By equation 24,

T_m(4-pole design) = 73.8 pound-feet

Other values are plotted in Figure 6.

It is obvious that with the increase in the number of poles, the winding factor cannot remain constant, and the appropriate terms must be changed. The nature of the error resulting can be illustrated by comparing the results on the 16-pole design in which the winding factor is assumed first to be 0.88 and finally to be unity. In the latter case,

$K_a=0.3198$
$K_c=1.50$ (unchanged)
$K_d=180.0$
$K_e=2.72$
$K_g=2.60$ (unchanged)
$K_h=38.0$
$K_m=11.8$

T_m (16 poles) = 18.2 pound-feet versus 16 by uncorrected winding factor

The use of the same lamination for all of these poles hardly would be practicable because of the fractional slot windings which might result, and the poor slot combinations and unfavorable harmonic conditions on some designs. The curve of inherent maximum torque is chiefly useful to observe the trend. It must be kept in mind, also, that all designs were made for the same tooth density, and that with an increase in numbers of poles the yoke densities decrease. As previously mentioned it is sometimes practicable, and reasonable, to increase the densities slightly with pole increase. This would result in raising the lower portion of the curve.

As this set of laminations was intended for frame 284 it is of interest to compare the maximum torque expected of standardized ratings (4 to 16 poles) in this NEMA frame, with that inherent with this lamination. While standardization requires a maximum torque of only 200 per cent of full load torque, this curve is plotted with a reserve, using 225 per cent ratio. These are the points connected by dotted lines in Figure 6. Between this reserve, and the flattening effect of the inherent torque curve brought about by flux density adjustment with pole change, inherent maximum torques can be kept safely above the values expected of the standardized ratings. Similar investigations on other diameters of laminations indicates the theoretical soundness of the ratings called for in standardized frames. However, the inherent torque reserve of some horsepower ratings over others is very marked in a few cases.

Reference

1. Some Aspects of Electric-Motor Design—Polyphase-Induction-Motor Design to Meet Fixed Specifications, T. C. Lloyd AIEE Transactions, volume 63, 1944, January section, pages 14–20.

Effects of Thermal Characteristics of Aircraft Generators on Load Analysis

STANLEY ROGERS
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Synopsis: The elaborate electric systems in modern long-range aircraft have made accuracy of load analysis more important than ever before. A more precise evaluation of the adequacy of proposed generating equipment can be made by averaging expected loads over several different periods which are chosen on the basis of the generators' thermal characteristics. These averages are compared graphically with curves which show the maximum overloads generators can tolerate without damage.

As AIRCRAFT have increased in size and complexity, so have their electric systems. The problem of establishing a precise relationship between system size and the anticipated electric loads has become more and more important.

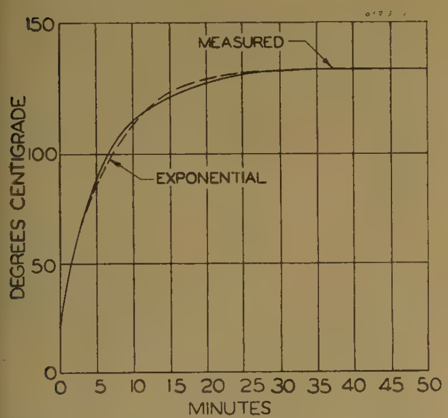

Figure 1. Simple exponential curve (broken line) compared with measured generator-winding heating curve (solid line)

Inadequate capacity may result in failure of the electric system with the consequent possible loss of the airplane, its crew, and its passengers. Too large a capacity, however, decreases pay load. Not only is extra fuel required to haul the useless weight, but additional fuel is needed to transport the fuel consumed in hauling this dead weight.

For airplanes with small electric systems, it may be sufficient to add up the power requirements of the various units of electric equipment and choose a generator which is estimated to be adequate. For somewhat larger systems, average power over various periods of time has been used as the basis for selecting generators. Later it was recognized by the more progressive manufacturers of large aircraft that peak loads and short time average loads are also important.

The most recent contributions to the making of an adequate load analysis include the following:

1. Loads should be averaged over not one or two, but several periods of time or "load intervals."

2. These periods of time depend on the overload thermal characteristics of the generator.

3. There is a maximum period of time over which loads may be averaged reasonably. The length of this maximum period depends on the generator.

4. A direct comparison between the maximum anticipated demand for electric power and the capacity of the generating system to meet that demand is essential to an adequate load analysis.

It is the purpose of this paper to explain why these propositions are true and to show how they are applied practically in determining the adequacy of an aircraft electric power system. The first step will be a consideration of the manner in which generator temperatures vary.

Temperature and Overloads

A typical aircraft generator is heated principally by the flow of electric current through its windings and is cooled by a blast of air under low pressure. The transfer of heat from the windings to the air follows the laws of thermodynamics, that is, the rise and fall of temperature occur exponentially. In most practical cases, the exact exponential function is complicated, but a simple function may be used as a first approximation. In Figure 1 a measured generator-winding heating curve (solid line) is compared with such a simple exponential curve (broken line). The close similarity between the two curves is apparent. In the analysis which follows it is assumed that the simple function is sufficiently accurate for practical purposes.

Experience has shown that a generator can stand a heavy overload for a short time without raising the temperature above a safe value. The less severe an overload is, the longer it can be tolerated without unwarranted damage. The length of time various overloads can be tolerated under a given set of operating conditions can be plotted. Figure 2 shows a typical curve. The curve is called a "permissible overload curve." The magnitude of overload (in terms of rated continuous-duty load) is plotted against the time for which the overload can be sustained without damage to the generator.

The permissible overload curve should be based on tests conducted by the manufacturer and should represent the guaranteed performance of the generator when

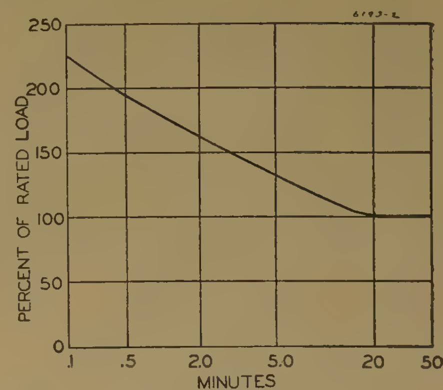

Figure 2. Typical permissible overload curve

subjected to overload. The curve rests on a specified set of conditions (generator speed, air pressure, air temperature, humidity, and so forth) and on assumed loading prior to the application of the overload. A family of such curves, representing performance under different conditions, would be even better. A conservative assumption is that thermal equilibrium under full-rated continuous-duty load has been attained before the overload is applied. Different conditions could be assumed, and curves based on them may be determined by test.

Unfortunately, manufacturers generally do not supply permissible overload curves of the type here suggested. It therefore has been necessary to develop a method for estimating them as accu-

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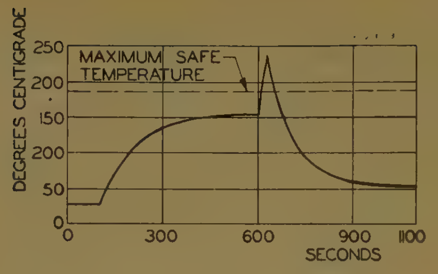
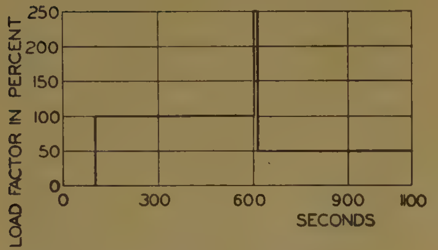
The approach utilized in the mathematical derivations in the appendix was suggested by Doctor George L. Shue, and the form and organization of the mathematical treatment is largely the work of B. D. Abramis, both of Consolidated Vultee Aircraft Corporation, San Diego, Calif.



A. Armature temperature curve



B. Corresponding load profile

Figure 3

The overload lasts five seconds. Load factor is the ratio of actual load to rated continuous-duty load

rately as possible. How this was done is set forth in detail in the appendix. Attention now is directed to the first of the four propositions which have been stated.

Why Loads Should Be Averaged Over Several Load Intervals

A number of load intervals are generally essential to a satisfactory analysis of loads on the generating system of a large airplane because averages taken over periods which differ substantially from the length of a damaging overload may give no indication of danger. This will be shown by means of two examples.

Figure 3A illustrates the variation of armature temperature with load. A cool generator is heated under rated load (as shown in Figure 3B) to full normal operating temperature. Thereupon a heavy overload is applied for five seconds, after which the load is reduced to one-half of full rated value. The attending rise and fall of temperature are shown in Figure 3A. It will be seen that during the overload the temperature quickly passes the maximum safe value and that about 25 seconds elapse before it decreases to a safe value. It will be noted that the graphs employ load factor as a parameter. It is the ratio of armature overload current (or power) to rated continuous-duty current (or power).

In Figure 4 the average load over intervals of 5, 15, 60, 180, and 600 seconds is plotted, the average always being taken so as to have the greatest possible value. The permissible overload curve is plotted to the same scale on this graph. The only average which exceeds the permissible overload curve is that for the 5-second interval. All the other averages lie on or below the curve and so do not indicate damage to the generator.

In Figures 5A, 5B, and 6, another case is illustrated in which a moderate overload is applied for 60 seconds. The armature reaches higher temperatures than in the first case and remains above the maximum safe value for more than a minute. This load is averaged over the same intervals of time as were used previously, and the averages again are compared with the permissible overload curve. Note that all the averages, except that for 60 seconds, lie below the permissible overload curve and give no warning of danger.

While the two cases chosen for purposes of illustration are not likely to be encountered in practice, experience has shown that when five well-chosen load intervals are used, excessive overloads may appear in only one or two of them, while the averages for the other load intervals appear to be safe.

From the foregoing considerations, it is evident that one or two periods of integration, regardless of their lengths, easily may fail to reveal the presence of damaging overloads. To discover with certainty every excessive overload, it would be necessary to use as many load intervals as there are different lengths of intermittent loads on the generating system; this would involve an impractical amount of work. Experience indicates that if the worst possible distribution of the intermittent loads is assumed, the use of four or five load intervals is usually adequate.

Load Intervals Chosen on the Basis of Permissible Overload Curve

If, several periods of integration must be used in order to make a load analysis which will reveal a dangerous overload, regardless of its duration, how long should each of the periods be? This question can be answered best by an example.

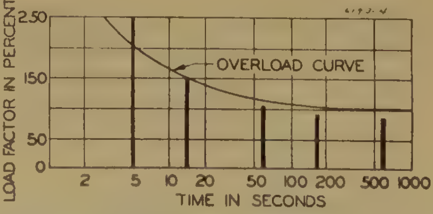


Figure 4. Values of load of Figure 3B averaged over periods of 5, 15, 60, 180, and 600 seconds

Figure 7 illustrates an overload characteristic curve for a large (40-kva) aircraft generator. Five periods over which loads may be averaged are indicated. These were chosen by the following criteria:

1. The shortest load interval should be somewhat longer than the shortest intermittent load worth considering and is usually of the order of five seconds. (Manufacturers sometimes supply overload ratings for such short times. It would be very helpful if a 5-second rating were furnished for all aircraft generators.)

2. The longest load interval should be equal to the shortest time for which the continuous-duty load ratings apply. Occasionally a manufacturer will supply this information. If he is unable to do so, the time may be estimated as being equal to three or four times the period of one cooling time constant.*

3. One of the intermediate load intervals should approximate the manufacturer's specified overload rating time (usually between one and five minutes).

4. The remaining intermediate load intervals should be so chosen as to make the differences between permissible overloads for successive intervals roughly equal.

5. The interval lengths chosen according to the foregoing rules should be modified enough to make the calculations easy.

It will be noted that the intervals indicated in Figure 7 are simply related to each other and that the overloads which correspond to them approximate an arithmetic progression as shown in Table I.

For one type of 6-kw aircraft generator, load intervals of 0.1, 0.5, 2.0, and 10 minutes are satisfactory. It is obvious that these load intervals would be unsatisfactory for use in analyzing loads on a 40-kva generator because of the marked difference in their thermal characteristics.

It follows, then, that before an adequate load analysis can be begun, the type of generator to be used first must be selected. Its thermal characteristics then are determined, and the load intervals are chosen. Loads may be averaged over each of these intervals, and the averages compared with the permissible overload (as indicated by the overload curve for the particular generator chosen). If the generators are not selected before the load intervals are chosen, a sufficient

*This time constant may be determined experimentally as follows: The generator is heated under standard conditions until it reaches thermal equilibrium under full rated continuous-duty load. With the machine rotating and the standard conditions maintained, the load is removed, and armature temperature is plotted against time as the machine cools. The cooling time constant is approximately the time required for the armature temperature to drop through the first 63.2 per cent of the difference between cooling air temperature and the armature's full load operating temperature.

number of different intervals must be analyzed to insure adequate coverage of the overload characteristics of all types of generators which might be used.

Maximum Time of Averaging Loads

Periods of one or two hours have been used in taking averages of loads on small generators. Some aircraft electrical engineers have been inclined to base all load intervals on the probable duration of each flight condition or aircraft operating condition (such as take-off, climb, cruise, land, and taxi) without regard to the thermal characteristics of the generators.

The duration of take-off, for example, is short and, for the analysis of loads during take-off, does justify limiting the longest load interval to a relatively short time. But it does not follow therefore

Table I. Load Intervals and Corresponding Loads for 40-Kva Aircraft Generator

Load Interval Length, Min	Permissible Percentage of Full-Rated Continuous-Duty Load	Difference in Percentages
0.1	222	
0.5	190	32
3	158	32
15	125	33
60	100	25

that, because an airplane may cruise for hours, the longest load interval for use in analyzing cruising loads should be one or more hours. On the contrary, the maximum length of a useful load interval is dependent, as will be shown, on the thermal characteristics of the generating equipment.

As the generator's heat is transferred to the cooling air, its temperature drops unless new heat is added to offset the losses to the air. The transfer to the air of the heat present in the generator at any particular moment takes place exponentially. The amount of heat available for transfer is limited by the temperature difference between the generator and the air. In the simplest case, 63.2 per cent of this heat will be transferred to the air after one time constant, 86.5 per cent after two time constants, 95.0 per cent after three time constants, and 98.2 per cent after four time constants.** Thus the cooling of the generator causes it to "forget" how hot it was at an earlier time.

Whenever the load on the generator changes, the temperature rises or falls exponentially. If the new load is maintained for three or four times constants, generator temperature will reach a value substantially independent of its previous magnitude.

The fact that the generator "forgets" how hot or cold it was is illustrated in Figure 8, in which four different load and temperature conditions are graphed. In each case the abnormal load ends at the point marked zero time constants. The restoration of normal operating temperatures proceeds exponentially. After three time constants the four curves are very close together, and after five time constants they merge. These curves show that

1. Temperature at any particular time is not affected materially by the temperature at any time more than three or four time constants earlier.

2. The temperature at a particular instant depends on those loads and only those loads which occurred during the immediately preceding three or four time constants.

3. Averaging loads over a period greater than three or four time constants may result in hiding a damaging overload by averaging it with an underload so far removed in time that it could have no effect on temperatures reached during the overload.

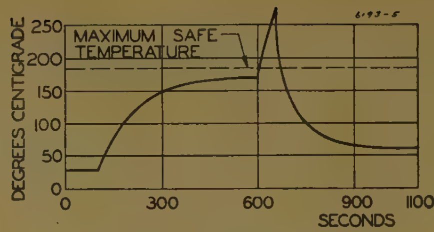
Choice of RMS or Average Values of Loads

A comparatively recent development in electric load analysis is the proposed use of rms values of load in place of average values. This proposal is based on excellent theoretical foundations. Since the load on aircraft generators is variable, its mean heating value is proportional to the average of the current squared. The suuare root of this value is a truer index of generator heating than is the average current. It has been found, however, that there is no significant difference between rms and average values in most practical cases. Since the rms values are much harder to calculate than average values, they rarely are justified in practice.

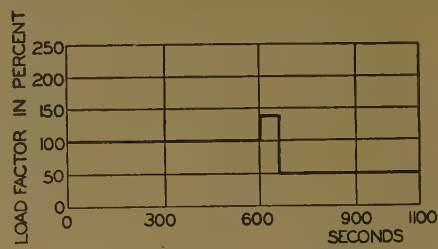
Direct Comparison of Anticipated Loads and System Capacity

The permissibe overload curve indicates the magnitude of overload which, under specified conditions, may be applied for a given time without damaging the generator. A direct comparison of the

**In the expression $(1-\epsilon^{-t/B})$, where t represents time, B is the time constant. This expression states the portion of the transferrable heat which has been conveyed to the air in time t.



A. Armature temperature curve



B. Corresponding load profile

Figure 5

Overload lasts 60 seconds

curve with the greatest anticipated average (or rms) loads for each load interval is obtained by plotting the calculated averages to scale on the graph of the permissible overload curve.

Such a graphical comparison is illustrated in Figure 9. The average loads appear as vertical bars. The height of each bar is proportional to the average value of the load it represents. The location of the bar on the time axis of the graph indicates the length of the period over which the average was taken. One graph is made for each flight condition (take-off, cruise, and so forth) that is analyzed.

A glance at the graph indicates whether there are any dangerous overloads, and, if there are none, what margin of safety the chosen generator provides. By examining a set of graphs for the most important flight conditions, those conditions which merit special study may be chosen. These graphs convey a great deal of information in simple easily-understood form.

Load Analysis Procedure

After the type of generator to be used has been selected tentatively, a curve representing its thermal characteristics on overload is obtained or calculated. Such a curve is shown in Figure 9. This curve is based on the assumption that the generator is heated fully under full load before the overload is applied. The curve shows the time for which any given steady overload then may be applied without causing excessive damage to the insula-

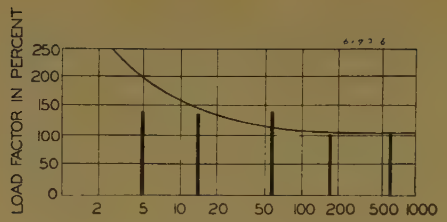


Figure 6. Values of load of Figure 5B averaged over periods of 5, 15, 60, 180, and 600 seconds

tion. The foregoing assumption is conservative since the chances are that, before and after an overload, the load on the generator is less than full rated capacity.

From the permissible overload curve four or five periods of integration or load intervals are chosen. These vary from a few seconds to 15 minutes or longer. They are selected in order to cover the range of permissible overloads in fairly even steps such as 100, 75, 50, and 25 per cent overload and no overload (full rated continuous-duty load). For convenience the lengths of the load intervals are made simple multiples or factors of each other.

The flight conditions or airplane operating conditions which are to be analyzed then are chosen. The most important ones are those in which the most severe electrical loads occur. These generally include take-off, cruise maximum, heavy combat, land, and ground operation.

Loads are tabulated and averaged for each flight condition and each load interval. The averages are plotted to scale as vertical bars on graphs of the overload characteristic of the generator. The resulting load analysis graph is illustrated in Figure 9. This graph not only summarizes the most important data, which conventionally are presented in tabular form on the load analysis chart, but also shows the relationship between calculated loads and available generator capacity.

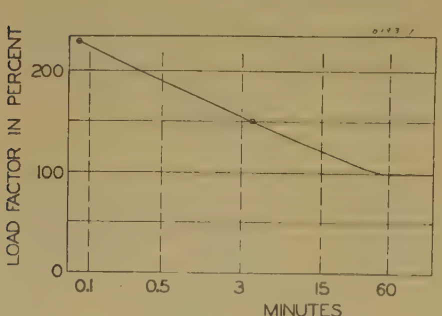


Figure 7. Permissible overload curve for 40-kva aircraft generator (calculated by method described in the appendix)

Five load intervals suitable for analyzing loads on this generator are indicated

While the propositions and method explained were applied to the analysis of loads on aircraft generators, the same principles may be used in analyzing many other types of electrical loads.

Appendix. Estimating the Permissible Overload Curve

Because manufacturers at present generally do not supply curves showing the time for which various overloads may be applied safely to the generator under specified conditions, a mathematical analysis of the thermal behavior of overloaded aircraft generators was undertaken.

The following simplifying assumptions have been made:

1. Heating and cooling occur as simple functions.
2. The armature heats and cools uniformly, and temperature gradients usually may be disregarded.
3. The highest temperatures occur in the armature winding.

An exact evaluation of the effect of field heating on armature temperature and of radiation and metallic conduction losses is not attempted.

The analysis made in this paper is necessarily approximate. The goal toward which it is directed is the expression of the time for which a given overload may be tolerated in terms of the load factor L (the ratio of actual load to rated continuous load) and a set of constants which can be evaluated. The analysis is divided into five operations.

1. Solution of the differential equation for rate of temperature change.
2. Expression of the temperature T in terms of the load factor L and time t.
3. Obtaining the expression for the permissible duration of a given overload.
4. Evaluating the ratio of the effective thermal capacity C to the coefficient of cooling K and incorporating the result into the final equation for permissible overload duration.
5. Drawing the permissible overload curve.

All temperatures are expressed in degrees absolute. The symbols and units used in the course of the analysis are explained as they occur.

Analysis of the thermal behavior of the armature winding is approached best by considering the rate of change of temperature. The truth of the following statement is evident:

$$\text{Rate of change of temperature} = \frac{\text{rate of heating} - \text{rate of cooling}}{\text{effective thermal capacity}}$$

Stated in mathematical terms, this becomes

$$\frac{dT}{dt} = \frac{HT - K(T - T_a)}{C} \tag{1}$$

where

T = Kelvin temperature of the armature winding

t = time in seconds

H = rate of heating per degree absolute and equals I^2R'/T' watts per degree Kelvin

I = armature current in amperes

R' = armature resistance in ohms at reference temperature T'

K = coefficient of cooling (watts per degree Kelvin)

T_a = Kelvin temperature of the cooling air

C = effective thermal capacity (joules per degree Kelvin)

This equation is based on the following reasoning:

1. For permissible generator temperatures the resistance R of copper varies roughly as its absolute temperature, $R = R'(T/T')$. Hence the rate of liberation of heat is proportional not only to the current squared, but also to the absolute temperature of the copper, $I^2R = (I^2R'/T')T$. The rate of heating is, therefore, HT.

2. The rate of cooling follows Newton's law of cooling and is proportional to the difference in temperature between the winding and the cooling air. It is represented by the expression $K(T-T_a)$ in which K is the factor of proportionality. This factor is presumed to be constant for a given installation under a given set of cooling conditions and is independent of H.

3. The rate of change of temperature varies inversely as the effective thermal capacity C.

Inasmuch as it is difficult to evaluate the constant K in practical cases, a solution in terms of temperature and load is desirable. The analysis will proceed as follows:

1. The differential equation 1 is solved (equations 2 and 3).

2. In the following steps (through equation 11), the ratio H/K is expressed in terms of the load and known temperature constants.

3. After expressing the temperature equation in a simplified parametric form (equation 13), the equation is solved for time t' (equation 14). The new equation contains a factor C/K which remains to be evaluated. This is accomplished by substituting two known boundary conditions in the overload time expression and solving the resultant equations simultaneously.

Operation I—Solution of the Differential Equation

$$\frac{dT}{HT - KT + KT_a} = \frac{dt}{C} \tag{2}$$

The solution of this equation is

$$T = \left(T_o + \frac{KT_a}{H-K}\right)\epsilon^{(H-K)t/C} - \frac{KT_a}{H-K} \tag{3}$$

where T_o is the initial temperature.

Note, that when $H = K$, equation 2 becomes

$$\frac{dT}{KT_a} = \frac{dt}{C} \tag{2a}$$

the solution of which is

$$t = \frac{C}{K} \times \frac{T - T_o}{T_a} \tag{3a}$$

Operation II—Expression of T in Terms of L and t

In the following steps equations 3 is expressed in terms of the ratio H/K. This ratio then is stated in terms of the load factor L and of S, a function of known temperature constants. The variable H thus is eliminated.

Letting $P = H/K$, equation 3 becomes

$$T = \left(T_o + \frac{T_a}{P-1}\right)\epsilon^{K(P-1)t/C} - \frac{T_a}{P-1} \tag{4}$$

When P is less than unity, that is, when H is less than K, the exponential term vanishes as t approaches infinity. This is clearly the case when the generator is carrying its full rated load. Letting the subscript n denote value at full rated continuous-duty load, the final steady state temperature is T_n

where

$$T_n = T_a/(1-S) \tag{5}$$

and

$$S = P_n = H_n/K = (T_n - T_a)/T_n \tag{6}$$

For overload conditions it may be assumed that H, the power consumed per degree absolute in heating the armature winding, is proportional to the square of the armature current. Letting L equal the ratio of load current to rated continuous-duty load current,

$$H = QL^2 \tag{7}$$

where Q is the factor of proportionality.

At full rated load, L equals 1, and H has its "normal" value of H_n. Therefore

$$Q = H_n \tag{8}$$

and

$$H = H_nL^2 \tag{9}$$

Substituting equation 6 in equation 9 and dividing each side by K,

$$H/K = H_nL^2/K = SL^2 = P \tag{10}$$

Substituting equation 10 in equation 4,

$$T = \left(T_o + \frac{T_a}{SL^2-1}\right)\epsilon^{(K/C)(SL^2-1)t} - \frac{T_a}{SL^2-1} \tag{11}$$

Let T_s be the asymptote of this equation. Its value is that of the constant term of equation 11.

$$T_s = \frac{T_a}{1-SL^2} \tag{12}$$

Using the parameter T_s, equation 11 becomes

$$T = T_s - (T_s - T_o)\epsilon^{-(K/C)(T_a/T_s)t} \tag{13}$$

Note, that for the special case in which $H = K$,

$$T = T_o + T_a(K/C)t \tag{13a}$$

Operation III—Solution for the Permissible Duration of an Overload

Solving equation 13 for time t when T equals T_p, the peak permissible winding temperature, and T_o equals T_n, the normal full rated continuous load (assuming that the generator is fully heated under rated continuous load before the overload is applied),

$$t' = \frac{C/K}{T_a/T_s} \ln \frac{T_s - T_n}{T_s - T_p} \tag{14}$$

For the case in which $H = K$,

$$t' = (C/K)(T_p - T_n)/T_a \tag{14a}$$

where t' is the permissible overload duration, that is, the time the given overload will require to raise armature temperature from its normal value T_n to the peak permissible value T_p, and T_s is the variable (see equation 12).

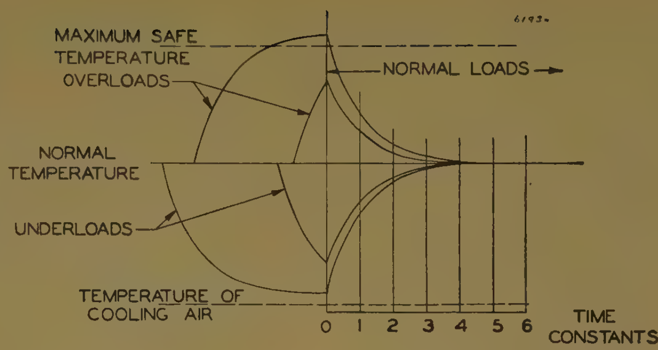


Figure 8. Temperature versus time curves showing that, regardless of deviations caused by underloads and overloads, temperatures return to normal a few time constants after normal load is restored (at time equals zero time constants)

When t' is plotted against the load factor L, it is seen that the curve is asymptotic to the line

$$L = \sqrt{(T_n/T_p)(T_p - T_a)/(T_n - T_a)}$$

Therefore equation 14 is useful only for load factors greater than this value. In practice typical usable values of L usually range from 1.5 to 2.5.

Operation IV—Evaluation of C/K and Statement of Final Solution

Attention now is directed to the quantity C/K. In the foregoing equations the thermal capacity C was assumed to be a constant. Actually, the effective thermal capacity is a function of both t and L. Taking into account the variation of C with time would have resulted in a more complicated differential equation and solution. Because of the uncertainty as to the nature of that variation, equations including this variation (some of which were set up and solved) were found to be of little value. However, treating C as a function of the load factor L only, does not affect the preceding analysis. The results obtained on this basis are in better agreement with experimental evidence than is an oversimplified solution in which the thermal capacity C is regarded as a constant.

The maximum value of C cannot exceed the thermal capacity of the material heated by the generator. There also must be a minimum value of C which holds only for a very brief period immediately following the application of a large overload. In so short a period there is no time for heat to be transferred from the copper of the armature winding to surrounding material. Because of variations in the resistance of different parts of the armature winding (a result of temperature differences for example) the effective minimum value of C may be far below the thermal capacity of the entire armature winding. The rate of heating at the hot spots is accelerated by the local increase in resistance caused by the increase in temperature.

Several assumptions as to the nature of the variation of C/K (K being a constant) were tried. The resultant permissible overload curves were plotted for a 40-kva aircraft generator for which two overload ratings were specified, one for five seconds and the other for five minutes. The following assumptions were evaluated from an examination of the curves:

1. C/K = constant.
2. $C/K = ML$.
3. $C/K = M/L + N$.
4. $C/K = M/(L + NL^2)$.
5. $\log C/K = N(M - L)$.

M and N represent constants, and L is the load factor. Only the last of these assumptions was found to be workable. Each of the other four either produced negative values of C/K for some range of values of L or failed to permit the permissible overload curve to pass through both specified rating points.

Letting t_q and t_r denote the durations of the two specified overloads and letting L_q and L_r respectively represent their load factors, by simultaneous equations we find

$$M = \frac{aL_r - bL_q}{a-b} \tag{15}$$

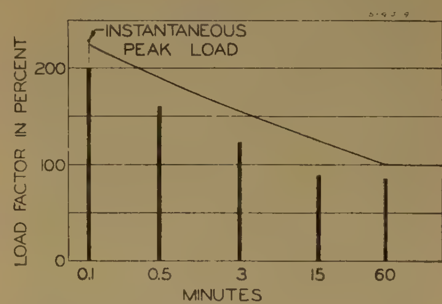


Figure 9. Typical load analysis graph

Bars represent average (or rms) values of load taken over periods of 0.1, 0.5, 3.0, 15, and 60 minutes

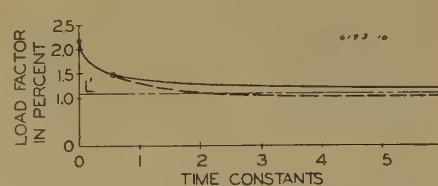


Figure 10. Illustration showing manner in which estimated (broken line) curve is joined to the calculated (solid line) permissible overload curve

The curve which is used consists of the estimated curve plus that portion of the calculated curve which lies to the left of the junction of the two curves

$$N = \frac{a-b}{L_r - L_q} \tag{16}$$

where

$a = \log (t_q/U_q)$
$b = \log (t_r/U_r)$

When $L = L_q$

$$U_q = \frac{T_s}{T_a} \ln \frac{T_s - T_n}{T_s - T_p} \tag{17}$$

When $L = L_r$

$$U_r = \frac{T_s}{T_a} \ln \frac{T_s - T_n}{T_s - T_p} \tag{18}$$

To evaluate equations 17 and 18, refer to equations 12 and 6.

Operation V—Drawing the Permissible Overload Curve

The permissible overload curve is calculated on the basis of known values for the following quantities:

T_a = absolute temperature of cooling air
T_n = normal hot spot insulation temperature (absolute) for rated continuous duty load
T_p = maximum permissible insulation temperature (absolute)
t_q = time for which an overload of load factor L_q can be applied to the generator and bring the hot spot insulation temperature exactly to T_p*
t_r = time for which an overload of load factor L_r can be applied to the generator and bring the hot spot insulation temperature exactly to T_p*

Figure 10 shows a calculated overload curve (solid line) and the two known overload rating points used in its determination. This curve is asymptotic to the line L-equals-L'. This is the magnitude of load which will just raise the temperature to T_p after an infinitely long time. Equation 14 indicates that this load can be carried indefinitely, but the generator manufacturer's rating for continuous duty is represented by the line L equals 100 per cent. For reasons already discussed, after three to four time constants have elapsed, the manufacturer's rating applies (refer to Figure 8). Hence the estimated curve (broken line) is added. It joins the calculated line below the lower of the two known overload rating points-and in the vicinity of three to four thermal time constants merges with the L-equals-100-per-cent line. It is recommended that the overload curve be plotted on semi-log paper and that the estimated curve be drawn in with a French curve.

Because of the composite nature of the permissible overload time curve, an inflection point may occur in the vicinity of the joint between the estimated and calculated sections. This point is especially likely to happen when the curve is plotted on semi-log paper

*In practice it may be assumed that t_q and t_r are the rated times for which overloads L_q and L_r may be applied.

Field Tests on Power-Line Carrier-Current Equipment

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Synopsis: This paper briefly describes a series of field tests made on the power line carrier current telephone system of the Pacific Gas and Electric Company, covering both amplitude and frequency modulated equipment. A discussion of various tests and test results is given, with a comparison of performance of the amplitude and frequency modulated equipment.

THE Pacific Gas and Electric Company has used carrier current telephone equipment on its high voltage transmission network for 23 years. Severe noise problems exist on some parts of the 230-kv transmission system because the lines have high corona losses. Early attempts to provide satisfactory carrier current communication over these lines were unsuccessful and led to the development of single-side-band power-line carrier-telephone equipment, some of which has been in service since 1927 and has been described previously.[1]

Major additions of hydroelectric generation and transmission facilities were made in 1944 requiring a considerable extension to the carrier current communication system. As some of the channels of the new carrier telephone system were to operate over the high corona lines, the noise problem required careful consideration. The use of frequency modulated carrier equipment[2] was proposed as a solution to the noise problem and an installation of this type equipment was made over a part of the system.[3] The complete carrier current telephone additions made in connection with the power system expansion in cluded four 2-frequency-duplex amplitude-modulated channels and one 2-frequency-duplex frequency-modulated channel on the Pacific Gas and Electric Company system. In addition, another 2-frequency-duplex amplitude-modulated channel was installed by the United States Bureau of Reclamation between Shasta Dam power house and the Pacific Gas and Electric Company Shasta substation. The carrier current telephone channels on the 230-kv system, with connecting channels, are shown in Figure 1.

The power system expansion was made to meet wartime requirements, and it was impossible to make complete tests of the carrier current equipment in the field at the time of installation because of th shortage of personnel, test equipment, an time. After the installation was con pleted, it was thought that maximum pe formance was not being realized from th system. Listening checks resulted in di ferences of opinion on the performance c various parts of the system. An extensiv investigation of the characteristics of th system was undertaken to obtain quanti tative data of the performance, and t make any adjustments found necessary t provide improvements.

Description of Tests

Data were taken to determine

1. The carrier frequency response of th equipment.
2. The audio frequency response.
3. The effects of transmitter output powe level on the received signal and signal-to noise ratio.
4. A comparison of frequency and ampli tude modulation signal-to-noise ratios.
5. The effects of percentage modulatior on signal-to-noise ratio.
6. Audio frequency attenuation of wir line and cable extensions used with th carrier current equipment.

The carrier frequency response gives ar indication of the correctness of the tuning of the transmitter output stages, the line tuning units, and the receiving equip ment. Carrier frequency response data were obtained by reading the transmitte output voltage, the coupling capacito current, carrier output current (measurec at the transmitter), received voltage, anc receiver limiter current as the transmitte oscillator frequency was varied. From

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The authors wish to acknowledge assistance in th tests given by R. W. Beckwith of the Genera Electric Company, Syracuse, N. Y., and L. L Creighton and A. K. Worthington of the Pacifi Gas and Electric Company, San Francisco, Cal.

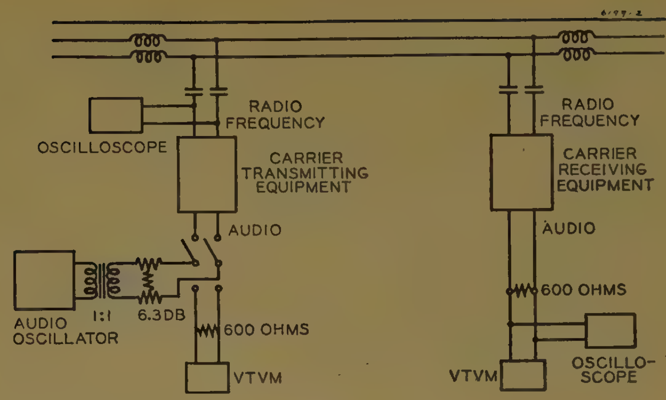

Figure 2 (above). Circuit used to obtain audio frequency response characteristics

Figure 1 (left). Carrier current telephone channels on the Pacific Gas and Electric Company 230-kv transmission system

these data it was possible to plot and observe the frequency response of each part of the equipment through which the carrier frequency signal passed.

Audio frequency response was obtained by supplying constant level audio frequency tones to the transmitter modulator and measuring the receiver output voltage. These tests were carried out after the correct carrier frequency adjustments had been established. The circuit used for the audio frequency response measurements is shown in Figure 2. These measurements were made on each channel of the system and over several channels in cascade. The input level of the audio tones was set to provide a high percentage modulation, approximately 75 per cent for normal voice, as determined by observation with an oscilloscope of the carrier frequency output. With this setting the modulation limiter would prevent overmodulation on peaks. Both the carrier frequency and audio outputs were observed with an oscilloscope to determine whether or not distortion was present.

Noise measurements were made on all channels and the signal-to-noise ratios were determined. Noise and signal measurements were made with a vacuum-tube voltmeter, and noise measurements are unweighted. Weighted noise measurements were taken on some channels, but an insuffiecent number are available for comparison and will not be presented.

Received signal and signal-to-noise ratios were observed as the transmitter power output was varied.

Frequency and amplitude modulation signal-to-noise ratios were compared on the same power transmission line section in order to obtain actual field data on the effectiveness of frequency modulation as a means of overcoming noise in carrier current telephone systems.

Audio frequency attenuation of the audio frequency extensions on the carrier system were obtained. As the extensions were of considerable length, their characteristics were considered to be a factor in over-all performance.

Results

It was found that considerable improvement was possible in some parts of the system, and data and curves showing the characteristics of some of the channels, both before and after readjustments, will be presented.

Carrier Frequency Response

Figure 3 shows the carrier frequency response of the frequency modulation channel with its original adjustment. These curves show a considerable dissymmetry about the 75-kc operating frequency, which was apparent as distortion in the receiver output. Careful retuning of the line tuning equipment resulted in doubling the transmission line current, a considerable increase in received voltage, and symmetrical variations about the operating frequency as indicated by the curves of Figure 4.

Audio Frequency Response

Considerable audio frequency response data were obtained. The manufacturer had guaranteed that the audio characteristics of the individual channels would be within plus or minus five decibels between 300 and 3,000 cycles. It was possible to come well within these limits on each channel. Audio characteristics of the 75-kc frequency modulation channel operating between Shasta and Contra Costa substations are shown in Figure 5. The audio response curve for the 85-kc amplitude modulation channel between Shasta substation and Pit 5 power house is illustrated in Figure 6. This curve is typical of the response of the amplitude modulation equipment.

One of the amplitude modulation re-

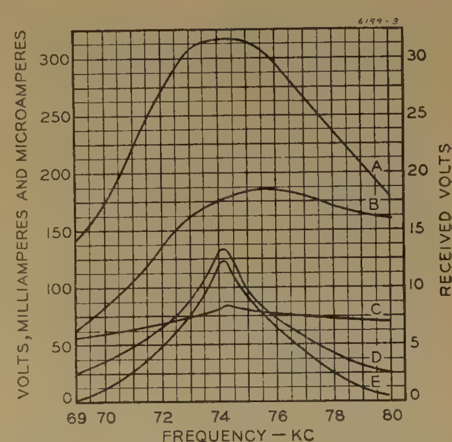

Figure 3. Carrier frequency response of 75-kc frequency modulation channel with improper line tuning

A—Output current at transmitter (milliamperes)
B—Transmitter output voltage
C—Coupling capacitor current (milliamperes)
D—Received voltage
E—Receiver limiter current (microamperes)

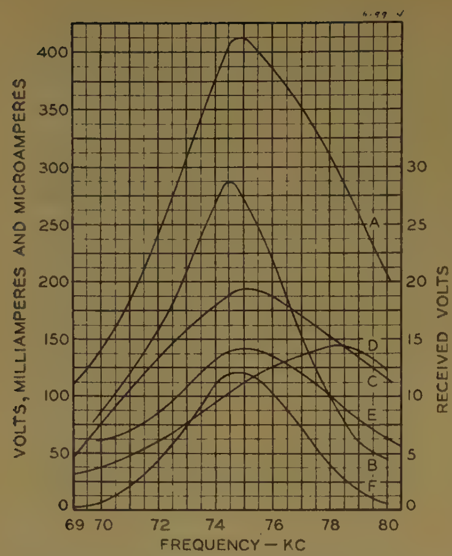

Figure 4. Carrier frequency response of 75-kc frequency modulation channel with correct line tuning

A—Output current at transmitter (milliamperes)
B—Received voltage (high *Q*)
C—Coupling capacitor current (milliamperes)
D—Transmitter output voltage
E—Received voltage (low *Q*)
F—Receiver limiter (high *Q*) current (microamperes)

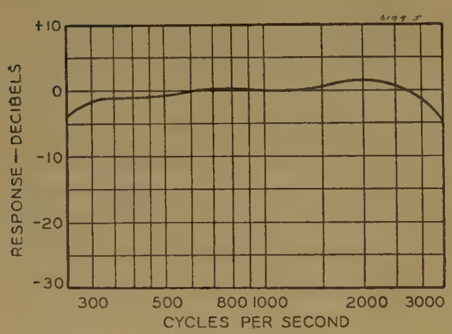

Figure 5. Audio frequency response of 75-kc frequency modulation channel

ceivers had been aligned for maximum detector current at the carrier frequency, and the effect on the audio characteristic is shown in curve *A* of Figure 7. When the intermediate frequency amplifiers were realigned for the correct band pass, the audio characteristic was improved considerably as indicated by curve *B* of Figure 7. At carrier frequencies the required band pass is a relatively greater percentage of the carrier frequency than in ordinary space radio applications. As in any highly selective receiver employing band pass circuits, tuning adjustments must be made carefully.

A limiting factor in audio frequency response, in this case of 2-frequency-duplex carrier system with interphase coupling using only 0.001-microfarad coupling capacitors, was found to be the resonance characteristics of the line tuning equipment. It was possible to improve the audio frequency response at the expense of received signal volts by reduction of the *Q* of the receiver line tuning equipment. Broadening of the audio frequency response resulted in some loss in signal-to-noise ratio. The effect on carrier frequency response is shown in the "low *Q* received volts" curve of Figure 4. An indication of the effect on the audio characteristics by broadening the band pass of the line tuning equipment is shown in Figure 8. Curve *A* shows the audio characteristics with high *Q* line tuning elements on the receiver input. Curve *B* shows the increased response to high frequencies with broadened receiver line tuning.

Over-all audio characteristics of three channels in cascade are shown in Figure 9. This system includes one frequency modulation and two amplitude modulation channels between station *X*, Oakland, and Pit 5 power house. The over-all response nearly meets the plus or minus 5-decibel limits set for one channel.

Power Output

The effects of the variation of transmitter output on the received signal and signal-to-noise ratio for the 75-kc frequency modulation channel are given in Table I. Received volts in Table I are signal plus noise. These results indicate the effectiveness of the frequency modulation receiver in maintaining the signal-to-noise ratio with large variations in the transmitter output and the received signal. It was possible to maintain communication over a 175-mile 230-kv line when using only the exciter unit of the frequency modulation transmitter with a power output of two watts.

Frequency and Amplitude Modulation Signal-to-Noise Comparison

A comparison of frequency and amplitude modulation signal-to-noise ratios is given in Table II. From these data it is apparent that the frequency modulation equipment is capable of maintaining a high signal-to-noise ratio even as the carrier signal approaches the noise level and consequently communication can be maintained under conditions when a conventional amplitude modulation system would be inoperative.

Table I

Transmitter Output, Volts	Receiver Input, Volts	Signal-to-Noise Ratio, Decibels
50	3.6	29.5
25	2.2	27.8
10.8	1.6	24.5
4.8	1.5	24.5

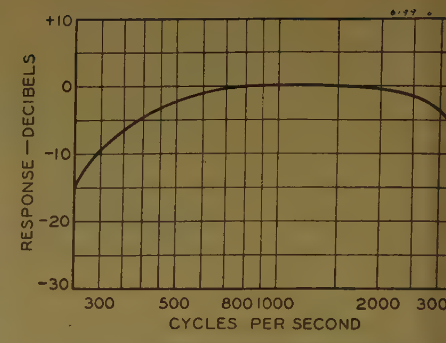

Figure 6. Audio frequency response o 85-kc amplitude modulation channel

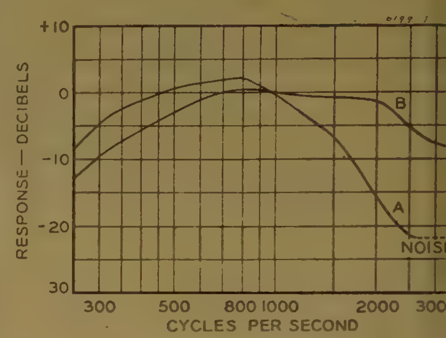

Figure 7. Audio frequency response of 140 kc amplitude modulation channel

Curve *A*—Receiver aligned for maximum de tector current at carrier frequency
Curve *B*—Receiver aligned for correct band pass

Effects of Modulation Level on Signal-to-Noise Ratio

On the frequency modulation channels with a deviation ratio of less than one-to-one, the signal-to-noise ratio was reduced considerably. Checks were made with deviation ratios of approximately one-third to one, and one-to-one, with the result that the signal-to-noise ratio was increased from 22 decibels with the lower to 34.5 decibels with the higher level of modulation. With a small deviation ratio, the phase modulation resulting from noise becomes appreciable relative to the desired frequency modulation and the noise output is proportionally higher. With the modulation correctly adjusted for normal voice input, overswing on peaks is limited by means of the modulation limiter which tends to maintain the deviation ratio near the correct value. A discussion of deviation ratios band-width requirements, and noise considerations is given in reference 2. One o

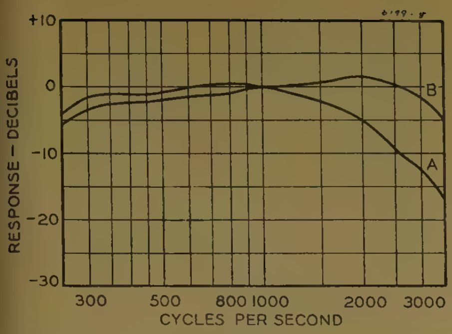

Figure 8. Audio frequency response curves showing the effect of the band pass of line tuning equipment

Curve *A*—High *Q* line tuning equipment
Curve *B*—Low *Q* line tuning equipment

Table II

Transmitter Output, Watts	Frequency Modulation Signal-to-Noise, Decibels	Amplitude Modulation Signal-to-Noise, Decibels	Frequency Modulation Over Amplitude Modulation Decibels
10	36.2	22.5	13.7
5	36	22.5	13.5
1	32.4	22.4	10.0
0.2	24.8	17.5	7.3

the early difficulties in the field adjustment of the frequency modulated carrier equipment was lack of a simple field check of deviation. With amplitude modulation equipment it is possible to observe the modulation envelope with an oscilloscope to check percentage modulation and the presence of distortion. A similar procedure for the frequency modulation equipment was suggested by R. W. Beckwith. With the carrier output connected to an oscilloscope with internal synchronization, a change of frequency will cause the pattern to move horizontally on the screen. By calibrating the horizontal movement versus frequency swing, a simple but sufficiently accurate method of measuring deviation is obtained, and the performance of the frequency modulation transmitter can be observed conveniently.

Audio Characteristics of Wire and Cable Extensions

The carrier terminals are located at high voltage substations at a considerable distance from the dispatcher and general offices of the company. The carrier telephone system was installed primarily to provide communication for the dispatchers to the generating plants, but also is used by others in carrying on business incidental to the operation of the power system. It was recognized that an investigation of the carrier system should include the characteristics of the extensions because they are a part of the complete communication system. The extensions to the dispatcher's office includes both open wire line and cable. Audio frequency characteristics of one of these extensions are shown in curve *A* of Figure 10. The extension from the dispatcher's office to the general offices is by submarine cable across the San Francisco Bay. The audio frequency characteristics of this cable are shown in curve B of Figure 10. There is in this cable a rapid cutoff of the frequencies above 2,500 cycles. The measured attenuations at 1,000 cycles were approximately two decibels for curve *A* and nine decibels for curve *B*, but the curves have been plotted through zero at 1,000 cycles for a better comparison of their shapes.

Listening tests also were made to compare various types of telephone instruments, and although no quantitative data were obtained, it was apparent that there was a large difference in the performance of different instruments, and that improvements could be made in some cases by providing modern high quality telephone instruments for use with the carrier equipment.

Conclusions

Tests such as those described in this paper provide information to evaluate properly the performance of the various parts of a carrier telephone system, and to make adjustments for optimum performance. Specifically it can be concluded that

1. Care in the adjustment of carrier current communication equipment will provide the maximum in performance, and adjustments cannot be made correctly without the use of proper instruments and a logical procedure.

2. Good audio frequency response is possible by correct adjustment of the equipment, and is particularly important where several sets are to be cascaded by means of audio coupling. Audio frequency response may be limited by characteristics of the line tuning equipment. In practice, a compromise between audio frequency response and signal-to-noise ratio must be accepted.

3. Frequency modulation equipment is capable of maintaining a high signal-to-noise ratio with large variations of carrier signal.

4. The frequency modulation system provides a definite advantage over amplitude modulation equipment insofar as signal-to-noise ratio is concerned, which permits successful carrier current operation on lines with noise levels so high as to preclude the use of conventional amplitude modulated equipment.

5. In frequency modulation, as in any carrier system, it is important to maintain a high level of modulation in order to realize the maximum signal-to-noise ratio.

6. Audio extensions from the carrier terminal equipment enter into the over-all consideration of the system, and can be limiting factors in the performance of a system.

Data such as those collected in the tests described are also invaluable in maintenance work on the system. The actual performance of the equipment can be compared with the test results from time to time, and any marked deviations can serve as guides in finding and correcting troubles.

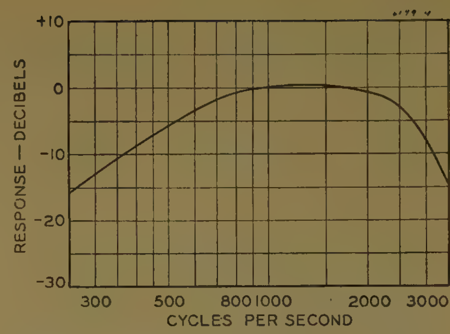

Figure 9. Over-all audio response of three carrier current channels in cascade

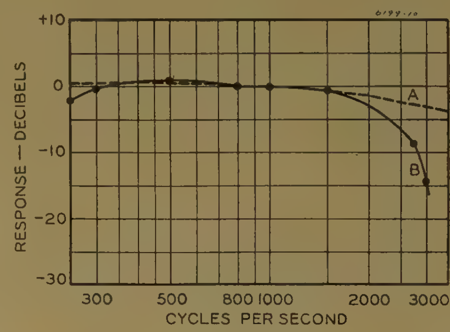

Figure 10. Characteristics of audio frequency extensions to carrier current channels

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Teaching Electricity and Magnetism

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FELLOW AIEE

Synopsis: A discussion of a method for presenting the fundamental relations of the science of electricity and magnetism is presented including a brief description of the various absolute unit systems. Most of the fundamental relations of electricity and magnetism are presented in general form. The relations are called general in the sense that they are complete and correct for use with any consistent unrationalized unit system. The relations are presented in a logical and rigorous sequence, that is, the basis of each relation is found in the preceding relations.

THE fundamental relations of electricity and magnetism form the basis of analysis for all analytical problems of electrical circuits and equipment. The primary objective of the formal training in engineering is to develop skill in engineering analysis. Thus it is very important that the electrical engineer have excellent training in his major basis for analysis, electricity and magnetism. Skill in engineering analysis demands an understanding of the field of application and limitations of basic relations which can be gained only through a well-organized study of the science involved.

The purpose of this paper is to present the author's approach to such study and training in electricity and magnetism. The organization presented has evolved through many revisions while presenting the material to sophomore electrical engineering students. It was developed on the basis of maintaining sequential rigor at the sophomore level of comprehension. Electrostatic field theory is developed toward the end of the organization instead of at the beginning because the average student has little background in this field and can gain essential background through the study of circuits and magnetic fields. For example, in accordance with the above principles potential difference is defined as work per unit charge instead of the line integral of field intensity.

Experimental, Defined, and Derived Relations, and Generalizations

The classification of the relations or formulas of electricity and magnetism into experimental, defined, or derived relations and generalizations was suggested to the writer by Professor Edward Bennett and forms an essential part of the organization of the text of which Professor Bennett is the coauthor.[1]

To comprehend fully the significance and limitations of each of the relations, formulas, and rules which make up the body of a science it is necessary to understand clearly the basis upon which each was obtained. All of the relations of electricity and magnetism can be classified under one or another of the four headings. The headings are so descriptive as to hardly need further exposition.

Experimental relations are those which are determined directly by experiment, such as Coulomb's law of force between charges and Ampere's law of force between parallel wires. Relations which are termed experimental were usually so determined historically but this does not imply that all relations which were originally determined experimentally should be so termed. For example, Joule's law concerning power dissipation in a resistor is considered derived in this presentation. It is also important to understand the actual experimental background for experimental relations because they must not be extrapolated too far beyond the experimental evidence on which they are based. Students of electronics are well aware of the limitations of $F = ma$ in cathode ray tube theory. For a presentation of original experimental procedures see Magie[2], and for a discussion of the limitations of experimental relations see Bridgman.[3]

Defined relations include all those which define units, conventions, and physical quantities. It is especially important to understand which quantities or relations are defined because a defined quantity must not be given properties beyond those involved in the definition. Failure to understand that the magnetic field is a defined quantity often leads to fruitless discussions concerning the mechanism of induced voltage. Also it is important that a given quantity be defined in one way only lest inconsistencies appear in succeeding analysis.

Derived relations are determined analytically usually through the combining of experimental and defined relations. Again familiarity with their source is important because all derived relations are circumscribed by the limitations of the experimental and defined relations upon which they are based.

Generalizations or postulates are relations which are assumed to be universally true because no exception to them has ever been observed. The law of conservation of energy is an example. Here again the engineer must understand that the law of conservation of energy based on the Macro measurements of engineering cannot be extrapolated into the realm of atomic physics.

Absolute Measure

A clear understanding of the origin and structure of unit systems is essential in engineering analysis. All electrical unit systems are based on absolute measure, that is, measurement based on the fundamental units of length, mass, time, and a fourth arbitrarily assigned quantity called the permeability of free space. Gauss laid the foundation for absolute measure in the 1830's by measuring the earth's magnetic field in terms of the moment of inertia and period of oscillation of one magnetized needle and of the deflection of a second magnetized needle in the vicinity of the first.[2] The pole strengths of both magnets cancel out of the equations thus giving the strength of the earth's field in terms of length, mass, time, and the permeability of free space tacitly assumed as unity in Gauss' experiments. Thus it was possible to duplicate magnetic measurements without a standard magnetic needle. All electrical units are based on similar measurements. The actual experimental work is very complicated involving difficult adjustments and corrections, but the basic academic definitions of the processes are quite simple as will be shown later.

The CGS Electromagnetic System

The cgs electromagnetic system is based on the centimeter, gram, second, and the permeability of free space taken as unity. It is the standard unit system in the field of physics and is the basis of the practical system of electrical units.

The International or Practical System

The early telegraph engineers did not find the cgs electromagnetic units a con-

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venient size so they developed the practical system whose units differed from the cgs electromagnetic units by various multiples of ten. However, the practical system embraced only circuit phenomena and was not extended to cover mechanics, or magnetic or electrostatic field theory. This was perhaps the greatest source of misunderstanding and confusion in the study of electricity and magnetism since any analysis involving mechanical force or electric or magnetic fields and electric circuits involved borrowing units from other unit systems and resulted in innumerable multiplying factors. Also, no relation could be considered complete without a statement of the particular assortment of units for which it was designed. Giorgi provided the solution in 1903 by the invention of the mks unit system. The system was adopted as the international standard in 1935.

The MKS System

Giorgi's invention of the mks system consisted of extending the practical system into a complete absolute system embracing all of mechanics and electric and magnetic field theory and at the same time retaining all of the circuit units of the practical system. This was done by choosing units of length and mass of such size as to make the unit of mechanical work equal to the unit of work in the practical system (the joule) and by assigning a numerical value to the permeability of free space ($\mu_v = 10^{-7}$) which would result in a unit of current equivalent to the ampere. These ideas will be expanded in the process of setting up the fundamental relations.

General Electrodynamic Equations

The foregoing discussion indicates the possibility of writing a complete set of fundamental relations in electricity and magnetism with no need for supplementary statements concerning the appropriate units. Units of the mks system are inserted on the right of a relation and the answer is known to be in units of the same system. It may not be quite as evident, however, that these same expressions serve equally well in any other consistent unit system. That is, the identical relations apply equally well in the mks, cgs electromagnetic, and cgs electrostatic systems.

With this approach, beginning students show equal facility in solving problems in any of the three unit systems mentioned. Often one or the other of the cgs systems will handle a given problem with greater convenience than the mks system. Experience has demonstrated that students can use three or more systems interchangeably without confusion. However, it is imperative that only one system be used in a given problem or relation.

The Fundamental Electrical Quantities

Electric Charge

The development of the fundamental relations presented herein is based on the concept of the electron as a charged particle, that is, charge is considered real and field is considered a condition which is produced in the vicinity of charge. The reverse of this may be a more exact picture of nature but the charged particle concept seems to lend itself more readily to the development of the science of electricity and magnetism.

Forces on Electric Charge

The science of electricity and magnetism deals with forces between electric charges at rest and in motion. We know from experiment that electric charges exert forces upon each other because of their positions, their velocities, and their accelerations.[4] The first effect is the basis of electrostatics, the second of generated voltages and magnetic forces, and the third of induced or transformer voltages and radio transmission.

Electric Fields

Even though it is possible in many cases to calculate the forces of electric charge in terms of "action at a distance," that is in terms of the force of one charge upon another, it is almost always simpler and more effective to assume that one charge or group of charges produces a field which in turn exerts a force on other charges in the vicinity. Forces between electric charges resulting from their relative positions are calculated in terms of electrostatic fields, and forces between charges in motion are calculated in terms of magnetic fields. The transmission of radio signals is usually calculated in terms of the electromagnetic field, but the calculations also can be made in terms of "vector potentials" which make no use of the magnetic field concept. The defining relations for the various field quantities are presented in the following section.

The Fundamental Electrodynamic Relations and Quantities

The fundamental electrodynamic relations and quantities are herein developed in a rigorous sequence. The source of each relation whether experimental, defined, derived, or generalization is so indicated. Space limitations prevent giving more than an outline of each derivation, or of stating the complete significance of all terms of all of the relations. The prime purpose in presenting the sequence of quantities and relations is to illustrate the logical manner in which the relations of electricity and magnetism may be developed. The reader should refer to standard text books in the field for details of specific relations. Units given are of the mks unrationalized system.

Basic Mechanics

1. l—length, meters (defined)
2. m—mass, kilograms (defined)
3. t—time, seconds (defined)
4. v—velocity, meters per second (defined)
5. a—acceleration, meters per second per second (defined)
6. Newton's law of force

$F=ma$ (experimental)

7. F—force, newtons (defined)

A newton is that force which will give a mass of one kilogram an acceleration of one meter per second per second.

8. W—work, joules (newton-meter) (defined)

$W=Fl$

9. P—power, watts, (joules per second) (defined)

$P=W/t$

Electric Circuit Quantities

10. Ampere's law of force between parallel wires (experimental)

$$F=\frac{K\mu_r I_1 I_2 l}{d}$$

where μ_r is the relative permeability of the medium between the wires, I_1 and I_2 are the currents in the two wires, l is the length of the wires, d the distance between centers, and K is a constant which determines the unit system. To establish an unrationalized system

$K=2\mu_v$

where μ_v is called the absolute permeability of free space. It is really a number which determines the size of the electrical units.

11. μ_v—absolute permeability of free space, (defined, fundamental mks unit)

$\mu_v=10^{-7}$

12. $\mu_r\mu_v$—absolute permeability of a medium (defined)
13. I—electric current, ampere (defined)

Unit current is defined in terms of Am-

pere's law of force between parallel wires

$$F=\frac{2\mu_r\mu_v I_1 I_2 l}{d}$$

The ampere is that current which will produce 2×10^{-7} newtons of force per meter of length between two parallel wires spaced one meter apart in air.

14. Q—electric charge (defined)

When two bodies attract each other with a force other than that of gravitation or repel each other with no material connection, they are said to be charged electrically.

15. Relation between electric current and charge (experimental)

$$Q=It$$

Rowland showed in 1876 that electric current was time rate of redistribution of charge.

16. Q—charge, coulomb (defined)

One coulomb of charge passes a given point in a circuit each second when the current is one ampere.

17. Potential difference (defined)

Potential difference is the action which tends to set up a current in an electric circuit. The unit is defined in terms of work per unit charge.

18. E, V—potential difference, volts (defined)

A potential difference of one volt is said to exist between the points a and b if one joule of energy is required to carry one coulomb of charge from a to b. The point b is at a higher potential than a if positive work is done on a positive charge.

$$V_{ba}=\frac{W}{Q}$$
$$=-\int_a^b \frac{F}{Q}\cos(F, dl)dl$$

where V_{ba} is the potential of the point b with respect to the point a, F is the electric force on the test charge, and (F, dl) is the angle between the direction of the electric force and dl.

19. Resistance (defined)

Resistance is that property of a circuit which opposes current flow.

20. Ohms Law (experimental)

$$I=E/R$$

21. R—resistance, ohms (defined)

A circuit is said to have a resistance of one ohm if a potential difference of one volt sets up a current of one ampere in it.

22. Resistance of a conductor (experimental)

$$R=\frac{\rho l}{A}$$

where ρ is the resistivity of the conductor material, l is the length of the conductor, and A its cross-sectional area.

23. G—conductance, mhos (defined)

Conductance is defined as the reciprocal of resistance.

24. Power in an electric circuit (derived)

$$P=VI$$

This relation is derived from the definition of potential difference, and of current considered as time rate of redistribution of charge.

25. Power in electric circuits (derived)

$$P=I^2R$$
$$=V^2/R$$

Derived from (24) and Ohm's law.

26. Energy in electric circuits (derived)

$$W=VIt$$
$$=I^2Rt$$
$$=V^2t/R$$

27. Law of conservation of charge (generalization)

28. Law of conservation of energy (generalization)

29. Kirchhoff's laws (derived)

$$\Sigma E's=0$$
$$\Sigma I's=0$$

These relations are derived on the basis of conservation of energy and conservation of charge.

30. Resistances in series (derived)

$$R_t=R_1+R_2+\ldots.+R_n$$

31. Resistances in parallel (derived)

$$R_t=\frac{1}{\frac{1}{R_1}+\frac{1}{R_2}+\ldots.+\frac{1}{R_n}}$$

Magnetism

Considering Ampere's law (13) let us assume that the current I_1 produces a stress in space which exerts a force on the current I_2. This assumption is the basis of the concept of magnetism. Let us say that the stress sets up a quantity called flux Φ (magnetism) usually represented by lines, and that the number of lines per unit area is flux density B.

32. Flux density (defined)

$$F=BIl\sin(B, l)$$

Flux density, a vector quantity, is defined by this relation where (B, l) is the angle between the direction of the flux lines and the direction of the wire.

33. B—flux density, webers per square meter (defined)

The flux density of a field is said to be one weber per square meter if a wire carrying a current of one ampere and perpendicular to the field experiences a force of one newton per meter of length.

34. Φ—flux, webers (defined)

$$\Phi=BA$$

or

$$\Phi\int B\cos(B, n)dA$$

The total flux passing through a plane area is equal to the product of flux density B and the area A if the flux density is uniform and perpendicular to the area. In case of nonuniform fields the summation must be determined by integration. In the integral, (B, n) is the angle between the direction of flux density and the direction of the normal to the differential area dA.

35. Flux density about a long straight wire (derived)

$$B=\frac{2\mu_r\mu_v I}{x}$$

where x is the distance in meters from the center of the wire.

36. Field intensity (defined)

$$B=\mu_r\mu_v H$$

Field intensity, a vector quantity, is defined by this relation. It is the agency which produces flux density; that is, the flux density at any point in a field is proportional to and directed along the field intensity vector.

37. H—field intensity, pragilberts per meter (defined)

A field intensity of one pragilbert per meter sets up a flux density of 10^{-7} webers per square meter in free space.

38. Field intensity about a long straight wire (derived)

$$H=\frac{2I}{x}$$

39. Magnetomotive force (defined)

$$M=\int H\cos(H, dl)dl$$

Magnetomotive force is defined as the line integral of field intensity. It is the total magnetic action along a given path. (H, dl) is the angle between the direction of H and the differential length of path dl.

40. M—magnetomotive force, pragilberts (defined)

A magnetomotive force of one pragilbert will set up a field intensity of one pragilbert per meter in a path one meter long.

41. Magnetomotive force about a long

straight wire (derived)

$$M=\int_0^{2\pi d}\frac{2I}{x}\cos(H,\,dl)dl$$
$$=4\pi I$$

It can be shown that the result is independent of the path of integration.

42. Magnetomotive force of a coil (derived)

$$M=4\pi NI$$

N is the number of turns in the coil which is carrying the current I.

43. Field intensity resulting from a current element (generalization)

$$dH=\frac{I\sin(r,\,dl)dl}{x^2}$$

where dH is the field intensity set up at a distance x from a differential length of conductor dl which is carrying a current I.

This relation must be considered true because no contradictory evidence has been discovered.

44. Field intensity about a wire of finite length (derived)

$$H=\frac{I}{r}(\sin\alpha_1+\sin\alpha_2)$$

Set up (43) for particular case and integrate. x is the perpendicular distance from the point at which H is evaluated to the axis of the wire. α_1 and α_2 are the angles between the perpendicular and the two ends of the wire.

45. Field intensity on the axis of a concentrated circular coil

$$H=\frac{2\pi r^2NI}{(r^2+x^2)^{3/2}}$$

where r is the radius of the coil and x is the distance from the plane of the coil.

Note. (44) and (45) become expressions for flux density by multiplying by $\mu_r\mu_v$.

46. Total flux produced in a toroid of cross-sectional area A and mean length l (derived)

$$\Phi=\frac{4\pi NI\mu_r\mu_vA}{l}\text{ (approximately)}$$

Ratio of length to area must be large.

47. Reluctance (defined)

The reluctance of a magnetic path is defined as the ratio of the magnetomotive force across a path to the flux set up in the path.

48. $\mathcal{R}$—reluctance, pragilberts per weber (defined)

Definition follows from name of unit.

49. Reluctance of a path (derived)

$$\mathcal{R}=\frac{l}{\mu_r\mu_vA}$$

Generated and Induced Voltages

50. Electromagnetic induction (experimental)

Faraday discovered that a voltage is generated whenever a conductor is moved across a magnetic field or whenever there is a change in the number of flux lines linking a coil. The first is a case of forces between moving charges and the second of forces between accelerated charges.

51. Voltage generated by a moving conductor (derived)

$$V=Blv\sin(B,\,v)$$

The derivation is made in terms of an elementary generator and the law of conservation of energy. Wire is considered moving perpendicular to its axis v is the velocity of the conductor and (B,v) is the angle between the direction of the flux and the velocity.

52. Flux linkages, weber-turn (defined)

$$\lambda=N\Phi$$

53. Voltage generated by changing flux linkages (experimental)

$$e=N\frac{d\Phi}{dt}$$

Note. The lower case letter e is used in this and subsequent relations to indicate instantaneous values. Instantaneous current is indicated in a corresponding manner.

54. Inductance (defined)

Inductance is that property of a circuit which tends to prevent a change in current.

55. L—inductance, henries (defined)

$$L=e/(di/dt)$$

A circuit is said to have an inductance of one henry if current changing at the rate of one ampere per second induces an emf of one volt.

56. Inductance in terms of flux linkages per ampere (derived)

$$L=N(d\Phi/di)$$

57. Inductance in terms of circuit dimensions (derived)

$$L=\frac{4\pi\mu_r\mu_vN^2A}{l}\text{ (approximately)}$$

Derived from (56) and (46).

58. Inductance of two long straight wires.

Distance d between centers, wire radius r, and length of circuit l (derived)

$$L=l\left(4\mu_v\log_e\frac{d}{r}+\mu_r\mu_v\right)$$

59. L_m—mutual inductance (defined)

$$L_m=e_1/(di_2/dt)$$

60. k—coefficient of coupling of two coils (defined)

$$k=k_1k_2$$

where $k_1=\Phi_{12}/\Phi_1$ and $k_2=\Phi_{21}/\Phi_2$. Φ_{12} is the part of flux Φ_1 which also links coil 2 and Φ_{21} is the part of flux Φ_2 which also links coil 1.

61. Mutual inductance in terms of self inductance (derived)

$$L_m=k\sqrt{L_1L_2}$$

62. Energy stored in inductive circuits (derived)

$$W=\frac{LI^2}{2}$$
$$=\frac{N\Phi I}{2}$$

63. Energy stored per unit volume in magnetic fields (derived)

$$\frac{W}{v}=\frac{\mu_r\mu_vH^2}{8\pi}$$
$$=\frac{BH}{8\pi}$$
$$=\frac{B^2}{8\pi\mu_r\mu_v}$$

where v is the volume in cubic meters.

64. Pull of an electromagnet (derived)

$$F=\frac{B^2A}{8\pi\mu_v}$$

Electrostatics

The following relations are concerned with forces experienced by electric charges resulting from their relative positions.

65. Coulomb's law of force between charges (experimental)

$$F=\frac{Q_1Q_2}{\epsilon_r\epsilon_vx^2}$$

where ϵ_r is the relative permittivity of the medium between the charges, also known as the dielectric constant. ϵ_v is the absolute permittivity of free space or more exactly a number the magnitude of which is determined by the unit system. (In an

Table I

Unit System	Permeability of Free Space	Permittivity of Free Space
Mks unrationalized...	10^{-7}	..1.113×10^{-10}
Cgs electromagnetic...	1	..1.113×10^{-21}
Cgs electrostatic......	1.113×10^{-21}..	1
Mks subrationalized..	$4\pi10^{-7}$	..8.854×10^{-12}

electrostatic system it is chosen arbitrarily and determines the size of the electrical units.)

66. ϵ_v—permittivity of free space (experimentally determined constant)

$$\epsilon_v = 1.113 \times 10^{-10}$$

The value may be determined by placing a known charge Q_1 at a given distance from another known charge Q_2 and measuring the force.

67. $\mathcal{E}$—Field intensity (electrostatic) (defined)

$$\mathcal{E} = \frac{F}{Q}$$

Similar to the case of parallel wires, let us assume that one charge produces a condition in space which exerts a force on the other charge. The condition is called field intensity and is defined by the foregoing relation.

68. $\mathcal{E}$—field intensity, newtons per coulomb (defined)

Unit field intensity exerts a force of one newton per coulomb on a point charge.

69. Field intensity in terms of space rate of change of potential, volts per meter (derived)

$$\mathcal{E} \cos (\mathcal{E}, dl) = \frac{-dE}{dl}$$

This relation is derived from the defining relations of field intensity and potential difference.

70. Field intensity about a point charge (derived)

$$\mathcal{E} = \frac{Q}{\epsilon_r \epsilon_v x^2}$$

71. D—electrostatic displacement density or flux density, pracoulombs per square meter (defined)

$$D = \epsilon_r \epsilon_v \mathcal{E}$$

Displacement density is set up by field intensity and is defined by the foregoing relation.

72. ψ—electrostatic displacement or flux, pracoulombs (defined)

$$\psi = DA$$

or

$$\psi = \int D \cos (D, n) dA$$

where n is the direction of the normal to the surface. Also see 34.

73. Displacement from a point charge (derived)

$$\psi = 4\pi Q$$

74. Displacement from any charge (derived)

$$\psi = 4\pi Q$$

from Gauss' theorem.

75. Capacitor (defined)

If a potential difference is applied to two conductors separated by an insulator, charge is observed to flow from one plate to the other. This arrangement is called a capacitor.

76. C—capacitance, farads (defined)

$$C = Q/V$$

A capacitor is said to have a capacitance of one farad if one coulomb is transferred from one plate to the other by a potential difference of one volt.

77. Capacitance of a parallel plate capacitor (derived)

$$C = \frac{\epsilon_r \epsilon_v A}{4\pi x}$$

78. Capacitance of a concentric cylinder capacitor (derived)

$$C = \epsilon_r \epsilon_v l / 2 \log_e (r_2/r_1)$$

where l is the length of the cylinder, r_2 the radius of the outer cylinder, and r_1 the radius of the inner cylinder.

79. Energy stored in a capacitor (derived)

$$W = {}^1\!/_2 CV^2$$

80. Energy stored per unit volume in an electric field (derived)

$$\frac{W}{v} = \frac{\epsilon_r \epsilon_v \mathcal{E}^2}{8\pi} = \frac{D^2}{8\pi \epsilon_r \epsilon_v} = \frac{\mathcal{E} D}{8\pi}$$

81. Force between parallel plates (derived)

$$F = \frac{2\pi\sigma^2 A}{\epsilon_r \epsilon_v}$$

σ = charge density.

82. Capacitor current (derived)

$$i = c(de/dt)$$

83. Capacitor voltage (derived)

$$e = \frac{1}{C} \int i \, dt$$

Table II

Quantity	Replace	By
Permeability	μ_v	$\mu_{rv}/4\pi$
Permittivity	ϵ_v	$4\pi\epsilon_{rv}$
Magnetomotive force	M	$4\pi M_r$
Field intensity	H	$4\pi H_r$
Displacement density	D	$4\pi D_r$
Displacement	ψ	$4\pi\psi_r$
Reluctance	R	$4\pi R_r$

Values of Permeability and Permittivity

Table I gives the values of absolute permeability and absolute permittivity in the four common unit systems. In the cgs electrostatic system the permittivity is assigned the value unity by definition and the value for permeability is experimentally determined. The other three, being electromagnetic systems, have the values of permeability assigned arbitrarily and the permittivity values are experimentally determined.

Examples

The application of various unit systems to the fundamental relations may be illustrated by a few simple examples.

Example 1. Calculate the force per meter of length between two long straight wires spaced 10 centimeters apart and carrying 500 amperes.

Mks (unrationalized)

$$F = \frac{2\mu_r \mu_v I_1 I_2}{r} = \frac{2 \times 1 \times 10^{-7} \times 500 \times 500 \times 1}{0.1}$$

$$= 0.5 \text{ newton}$$

Cgs electromagnetic

$$F = \frac{2 \times 1 \times 1 \times 50 \times 50 \times 100}{10}$$

$$= 50{,}000 \text{ dynes}$$

Cgs electrostatic

$$F = \frac{2 \times 1 \times 1.113 \times 10^{-21} \times 500 \times 3 \times 10^9 \times 500 \times 3 \times 10^9}{10}$$

$$= 50{,}000 \text{ dynes}$$

The solutions check since one newton equals 10^5 dynes. The expression does not apply to the rationalized system.

Example 2. Calculate the flux density at 10 centimeters from a long straight wire carrying 50 amperes.

Mks unrationalized

$$F = \frac{2\mu_r \mu_v I}{r} = \frac{2 \times 1 \times 10^{-7} \times 50}{0.1}$$

$$= 1 \times 10^{-4} \text{ weber per square meter}$$

Cgs electromagnetic

$$F = \frac{2 \times 1 \times 1 \times 5}{10}$$

$$= 1 \text{ gauss}$$

$$= 1 \text{ maxwell per square centimeter}$$

The calculation can be made similarly for the electrostatic system but there is no

accepted name for magnetic flux density in the electrostatic system. The results check again since gausses $\times 10^{-4}$ equals webers per square meter.

Example 3. Calculate the capacitance of a parallel plate capacitor consisting of two plates 50 centimeters square and separated by one centimeter. Neglect fringing.

Mks unrationalized

$$C=\frac{\epsilon_r\epsilon_v A}{4\pi d}$$

$$=\frac{1\times 1.113\times 10^{-10}\times 0.25}{4\pi\times 0.01}$$

$$=2.22\times 10^{-10}\text{ farads}$$

Cgs electrostatic

$$C=\frac{1\times 1\times 50^2}{4\pi\times 1}$$

$$=198.8\text{ stat-farads}$$

Cgs electromagnetic

$$C=\frac{1\times 1.113\times 10^{-21}\times 50^2}{4\pi\times 1}$$

$$=2.22\times 10^{-19}\text{ abfarads}$$

See Hudson[5] for a complete list of conversion factors.

Fundamental Rationalized Relations

The fundamental relations can be changed to the rationalized form by making the substitutions indicated in Table II wherever any of the quantities listed appeared in the relations.

The question of the choice between the rationalized and unrationalized systems is not the subject of this paper.

Unit Magnetic Pole

The unit magnetic pole has not been used in connection with the organization because the approach to magnetism through forces on current carrying conductors seems more satisfactory to the writer. Too often the student is introduced to electrical and magnetic quantities through the concept of impossible experiments with unit poles. However, it is granted that the concept may be useful in certain field analyses and there is no objection to defining a unit pole for this purpose just as we have defined magnetic fields for the purpose of simplifying the calculation of forces between electric charges in motion.

Summary

Most of the fundamental relations in electricity and magnetism are presented in general form. The relations are called general in the sense that they are complete and correct for use with any consistent unrationalized unit system. The basis of each relation is clearly indicated and the relations appear in a logical sequence of development. That is, the basis of each succeeding relation is found in the preceeding relations. Such an organization presents the beginning student with the means of acquiring a fundamental understanding of the science of electricity and magnetism, and also forms his basic training in engineering analysis.

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150,000 Horsepower Applied to Aeronautical Research

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BEHIND THE SCENES of the rapid development of aeronautics in this country during recent years are the aeronautical research laboratories. In addition to their part in the development of aeronautics, these laboratories are of interest because of the variety of electric devices they employ and the surprising magnitude of their electric power loads. One of the largest aeronautical laboratories in terms of electric power is the Ames Aeronautical Laboratory of the National Advisory Committee for Aeronautics, located at Moffett Field, Cal. This laboratory, on which construction was started in 1939, has developed rapidly to the point where it is now the largest single connected load on the Pacific Gas and Electric Company's system. With completion of wind tunnels now under construction this laboratory will have over 150,000 connected horsepower.

The Ames Aeronautical Laboratory is one of three laboratories of the National Advisory Committee for Aeronautics (NACA) the other two laboratories being at Langley Field, Va., and Cleveland, Ohio. The National Advisory Committee for Aeronautics consists of 15 members appointed by the President and serving without pay. This committee serves in an advisory capacity and is charged with the responsibility of studying the fundamental problems of flight with a view to their practical solution. A large supporting staff of full-time employees under the director of aeronautical research operates the facilities just mentioned, and carries out all types of appropriate basic studies of interest to not only the military services and the industry but also to the air lines and private flyers. The Ames laboratory includes several of this country's most up-to-date and highest powered wind tunnels, as well as a highly developed flight research department.

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The author acknowledges the assistance of the General Electric Company, the Westinghouse Electric Corporation, and the Allis-Chalmers Manufacturing Company by supplying the detail design of the various drive systems and for their ingenuity and co-operativeness in recommending suitable types of equipment to meet the requirements of the Ames Aeronautical Laboratory.

Wind Tunnels

The largest users of power at the Ames laboratory are the wind tunnel fan drives. From the standpoint of control requirements and novel applications of electric machines they are also among the most interesting. A wind tunnel is simply a device for circulating air past an airplane or model of an airplane to simulate the conditions which occur in flight. A schematic sketch of a typical wind tunnel is shown in Figure 1. The propeller, or fan, circulates air around the continuous closed air passage in the direction shown. The air speed is increased to a maximum at the test section, where the model is

mounted, by the nozzle-like shape of the air passage at the approach to the test section. Beyond the test section the passage gradually is expanded again to reduce the air velocity and minimize friction losses.

The airplane or model undergoing tests is supported on a balance system so that the forces of the wind on the model may be measured to determine its lift, drag (resistance to forward motion), pitching moments, and so forth. In addition to the over-all forces on the model, forces on individual components may be measured by suitable techniques such as pressure-distribution measurements or by special measuring devices built into the model. The wind tunnel provides a simple device in which controlled test conditions are readily obtainable, and the effects of changes to the model may be studied quickly, accurately, and without the hazards of experimental flight testing. Figure 2 shows a powered model mounted in the 7-by-10-foot wind tunnel.

The power required to operate a wind tunnel ranges from a few horsepower up to many thousands of horsepower, depending upon the size and the air speed produced. The ideal wind tunnel would test full size airplanes at maximum flying speeds. However, such a wind tunnel would require over 400,000 horsepower; consequently, all wind tunnels to date have represented compromises, either as to size or maximum air speed, or both. In the early days of aeronautics much valuable information was obtained from small low speed wind tunnels of a few horsepower. As progress in aeronautics has demanded more and more attention to details the scale-effect errors of such small low speed tunnels no longer can be tolerated. Moreover, as the speed of flight approaches or exceeds the speed of sound (approximately 765 miles per hour at sea level) important phenomena caused by compressibility of the air occur which cannot be simulated in a wind tunnel except by operation at the same high speeds. Thus one finds at the Ames laboratory, for example, the following wind tunnels:

1. The 40-by-80-foot full scale wind tunnel with a test section 80 feet wide by 40 feet high capable of testing actual full size airplanes up to about 70-foot wing span at air speeds up to 250 miles per hour.

2. The 16-foot high speed wind tunnel (16-foot-diameter test section) which tests relatively large models up to about 90 per cent of the speed of sound.

3. The 1-by-3-foot supersonic wind tunnel which tests small models at speeds equivalent to 1,600 miles per hour at sea level.

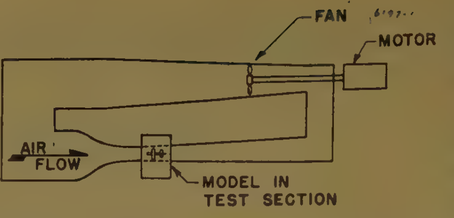

Figure 1. Schematic sketch of the essential features of a wind tunnel

These tunnels require, respectively, 36,000, 27,000, and 10,000 horsepower. Four other wind tunnels of various particular characteristics make a total connected wind-tunnel-fan horsepower of 87,950. An additional 50,000-horsepower supersonic wind tunnel now under construction, plus the miscellaneous laboratory equipment, will bring the total connected horsepower of the Ames laboratory to well over 150,000 horsepower. Figure 3 shows the 27,000-horsepower 16-foot high-speed wind tunnel.

The requirements for the drive system for a wind tunnel are quite exacting in that they usually require adjustable speed over a very wide range and steady speed at any given speed setting, while at the same time often requiring tremendous motor sizes. Operation is intermittent and the equipment must withstand frequent starting, stopping, and speed changing. In large sizes, power factor of the drive system may be important, and starting surges or surges during speed changing may be a critical factor for large wind tunnels.

Of the various types of adjustable speed drives available, the d-c adjustable voltage system (in which the propeller is driven by a d-c motor supplied by a separate generator, and the speed control is obtained by varying the voltage of the generator) has been almost ideal for these applications because of its simplicity of control, wide speed range without discontinuities, and adaptability to automatic control. This system has been used for most small and medium sized wind tunnels. Various other types of systems have been used for larger tunnels, either in an attempt to decrease the cost over that of the d-c adjustable-voltage system or to avoid problems which arise in applying the d-c system in very large sizes. In recent years what is sometimes called the modified Kraemer system has been used very successfully on several large wind tunnel drives, including the Ames 36,000- and 27,000-horsepower tunnels. In this system the main propeller motor is a wound-rotor induction motor, and speed control is obtained by impressing a voltage of controllable frequency on the rotor of the induction motor.

Modified Kraemer Adjustable Speed Drive

To understand the operation of the modified Kraemer system, it might be well to consider first the action of the wound-rotor induction motor with conventional speed control by means of secondary resistance. The voltages and power in the rotor circuit are indicated in Figure 4a. For ease in understanding, the rotor circuit is shown in Figure 4 as though single phase, and, in determining components of power, internal losses are neglected. As the resistance connected across the rotor is increased, the rotor must slip a greater amount in respect to the rotating stator flux in order to generate sufficient voltage in the rotor circuit. This circulates sufficient current in the rotor to develop the necessary torque, resulting in a decrease in speed. By varying the rotor resistance a range of speeds may be obtained. The current and voltage in the rotor circuit will be alternating at a frequency corresponding to the slip between the rotor and the stator flux. As far as the rotor circuit is concerned, the motor acts like a generator and, neglecting motor internal losses, induction motor theory shows that

Figure 2. A powered model mounted in the 7-by-10-foot wind tunnel at the NACA Ames Aeronautical Laboratory, Moffett Field, Calif.

an amount of power is dissipated in the resistor equal to Ps, where P is the power input to the motor and s is the slip in fraction of synchronous speed. Thus power is wasted, resulting in a low efficiency which cannot exceed $(P - Ps)/P = 1 - s =$ speed in fraction of synchronous speed.

If it were possible to generate a voltage whose frequency would at all times be equal to the slip frequency of the motor and in the proper phase relationship, it would be possible to substitute this generated voltage in place of the voltage drop across the resistor. An alternator may be considered as a source for such a voltage. At first glance it would seem impracticable, in replacing the resistance by an alternator, to keep the generated frequency the same as the rotor frequency so that the generated voltage at all times would be exactly opposite in phase to the voltage induced in the rotor circuit. However, upon studying the arrangement further, it will be found that any tendency for the alternator voltage to change in frequency or phase in respect to the rotor voltage will result in a circulating current which will produce a synchronizing torque tending to prevent the two from getting out of phase. This action is very similar to that which occurs in the case of a synchronous motor which, within the limits of its pull-out torque, always will remain with its internal generated voltage in phase opposition to the line voltage. If the frequency of the alternator is decreased by decreasing the alternator speed, the synchronizing current induced between the alternator and the induction motor rotor will increase the motor torque causing the rotor to speed up, thus decreasing the slip frequency and permitting the two to remain in step. Conversely, if the alternator frequency is increased, the resulting circulating current will cause the induction motor to decrease in speed so that the increased slip frequency still matches the alternator frequency. Thus, the speed of the induction motor may be controlled over the operating range by varying the speed of the alternator connected to the rotor. Similarly, as the main motor is loaded and tends to slow down, the corresponding phase shift between the rotor voltage and the alternator voltage causes a circulating current to flow which produces the necessary torque to support the load. The same circulating current flowing through the alternator tends to speed up the alternator, causing it to act as a motor. The flow of power is shown in Figure 4b.

Figure 3. The 27,000-horsepower 16-foot high-speed wind tunnel at the NACA Ames Aeronautical Laboratory

The speed of the alternator can be adjusted most readily by driving it by a d-c motor supplied from a d-c generator with adjustable voltage speed control. The complete system is outlined in Figure 5 for the case of a single fan motor.

The alternator must be of sufficient size to handle the full load rotor current of the induction motor, and must be able to produce, at the primary frequency, a voltage equal to the rotor voltage of the induction motor at standstill; although, with a fan load, the conditions of maximum current and maximum voltage never occur simultaneously. The power which must be handled by the motor-generator sets under any particular condition of load and speed must be equal to the power in the induction motor rotor circuit which otherwise would have been dissipated in a secondary resistance. This power, as previously stated, is equal to Ps, or, in terms of the motor output, the rotor power is $P_o[s/(1\text{-}s)]$, where P_o is the power output. For a fan load varying as the cube of the speed, such as is encountered in the usual wind tunnel drive, the rotor power reaches a maximum value at two-thirds synchronous speed equal to approximately one-seventh of the motor horsepower rating. Thus, it may be seen that most of the auxiliary machines are much smaller in size than would be the motor-generator set for a straight d-c adjustable-voltage system.

This modified Kraemer system is considerably more complicated in the number of machines involved and the complexity of the control than a straight d-c adjustable-voltage system. In fact, it involves, in addition to the wound-rotor motor, a complete adjustable-voltage system for secondary frequency control. The advantage of the modified Kraemer system on extra large drives lies in the reduced size of the d-c machines, and the fact that the equipment may be started with less disturbance to the power system than in the case of starting the adjustable-voltage d-c system. This system has essentially all the desirable control characteristics of the d-c adjustable-voltage system, including ready control of power factor and good efficiency.

In the usual starting sequence the constant speed motor–generator set is first started. This set is quite small compared with the rating of the main motor and offers no difficulties from the standpoint of starting surge. The variable speed set then is brought up to speed with the alternator excited until the alternator frequency is 60 cycles. Since

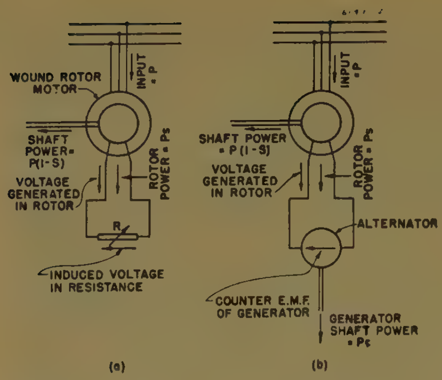

Figure 4. Voltage and power relations in the circuit of a wound-rotor induction motor

(a.) With resistance in secondary circuit
(b.) With counter electromotive force of an alternator in secondary circuit

the alternator is connected to the main motor rotor at all times, a voltage appears on the primary of the motor. By adjusting the alternator speed the motor stator voltage may be synchronized with the line voltage and the motor primary breaker closed with no attendant power disturbance. The motor then may be caused to rotate at any desired speed by reducing the speed of the alternator. With this system a 27,000-horsepower motor can be brought onto the line with no more disturbance than that incurred in starting a 5,000-horsepower synchronous motor.

While exciting the main motor from the secondary side prior to synchronizing, the d-c machine of the constant speed set is generating, driving the variable speed set. After synchronizing, when the alternator frequency is reduced and the main motor starts rotating and picking up load, the flow of power is reversed. Thus the various machines operate sometimes as generators and at other times as motors, and the terms generator or motor are meaningless.

Figure 5 shows a single fan motor. The modified Kraemer system functions equally well if the load is divided among two or more motors with their rotors connected to a single alternator. In fact the 36,000-horsepower wind tunnel, because of its tremendous size, actually uses six 6,000-horsepower motors and fans arranged aerodynamically in parallel in a double row of three above three (Figure 6). The motors are mounted in the air stream, housed in the faired enclosures immediately in front of the propellers. Since all motors are connected to a common frequency on the stator side and to another common frequency on the rotor side, they operate in exact synchronism with each other. In the 16-foot wind tunnel the 27,000 horsepower is divided between two 13,500-horsepower motors driving separate counterrotating fans in tandem (Figure 7). The two motors are mounted back to back in the steel enclosing drum between the two fans.

A doubly fed induction motor is similar to a synchronous motor in that it has a revolving stator field and a separately excited rotor field, the two fields being locked into step by the synchronizing torque. As a synchronous motor tends to oscillate, or hunt, so also does the doubly fed induction motor. A damper winding cannot be used on the induction motor, as on the synchronous motor, because the magnetic flux normally rotates in respect to the rotor at slip frequency and would generate a high voltage in the damper winding even though there were no hunting. Thus the hunting problem demands careful attention on the part of the designer. Where there is a single motor, damping can be incorporated in a damper winding in the variable-frequency alternator. However, if two or more induction motors are connected in parallel to a single alternator, it is possible for the motors to oscillate in respect to each other without affecting the alternator, thus obtaining no help from the damper winding. In the 16-foot wind tunnel with its two motors, the manufacturer avoided independent oscillations between the motors by connecting the variable frequency alternator in series in the rotor circuit between the two motors so that circulating current between the two motors also must pass through the alternator. In the 40-by-80-foot wind tunnel with its six motors there was no alternative to the parallel connection, and hunting was prevented only by careful attention to damping factors in the design.

Figure 8 shows the control motor–generator sets at the 16-foot wind tunnel.

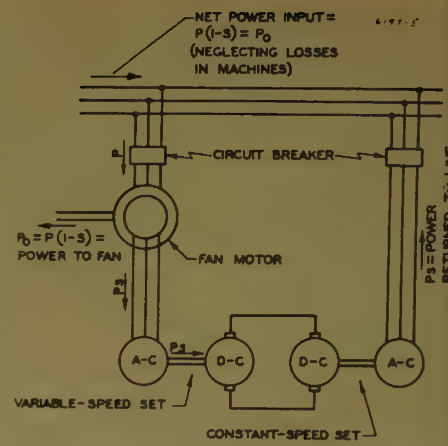

Figure 5. Connection diagram for a complete modified Kraemer system

Two d-c machines are seen on each set because it was impracticable to handle the full power in a single d-c machine. Figure 9 shows a view of the associated switchgear. Modern metal clad switchgear is used throughout, using oil circuit breakers of 500,000 kva interrupting capacity. Controls are semiautomatic so that the wind tunnel operator can devote his full attention to the test under progress. Only such simple controls as a "start-stop" switch and a speed control switch are required to give the operator complete control of the massive wind tunnel drive equipment. When starting a run the operator simply turns the start-stop switch to "start." Without further attention on his part the equipment automatically goes through the sequence of starting the various auxiliary machines and motor–generator sets and automatically synchronizing them with the line. When the main motors are on the line a red light is illuminated on the control panel signalling to the operator that the equipment is on the line ready for him to adjust the speed as needed. The complete start consumes approximately four minutes.

Figure 6. View looking upstream at fans of the NACA 40-by-80-foot wind tunnel—six fans, each driven by a direct - connected 6,000 - horsepower motor

Combination Slip-Regulator and D-C Drive

The 12-foot pressure wind tunnel, with 10,250 fan horsepower, uses to advantage a combination of d-c adjustable-voltage control and wound-rotor induction motor with secondary resistance control. The fan drive unit consists of a 1,250-horsepower d-c motor and a 9,000-horsepower wound-rotor induction motor in tandem. A motor–generator set supplies adjustable voltage direct current for the d-c motor, while a liquid rheostat provides an adjustable resistance for the induction motor rotor circuit. Since the fan load of a wind tunnel varies approximately as the cube of the speed, a tunnel requiring 10,250 horsepower at full speed would require only 1,280 horsepower at one-half speed. Thus the 1,250-horsepower d-c motor provides ample power up to almost half speed. Above half speed the induction motor is energized and its secondary resistance adjusted so that it carries the excess load. At full speed both motors are carrying full load. Sensitive automatic speed regulation is maintained at all speeds through the d-c motor to prevent fluctuations in speed.

Supersonic Tunnel Drives

The wind tunnels discussed thus far all operate at test air speeds below the speed of sound. Wind tunnels designed for operation at test section air speeds in excess of the speed of sound, that is, supersonic wind tunnels, exhibit some quite different characteristics.

1. In the supersonic tunnel the volume of air handled is relatively small and pressure differences relatively high; consequently, the "fan" of a supersonic tunnel has less of the characteristics of an ordinary fan or airplane propeller and is more like a turbine or air compressor. In fact, the 1-by-3-foot supersonic wind tunnel uses commercial centrifugal type air compressors to move the air.

2. At supersonic speeds, an attempt to increase the air speed by speeding up the blower results primarily in an increase in air density at the test section rather than an increase in air speed. Hence air speed control in this class of wind tunnel must be obtained by changing the nozzle contours and the cross sectional area of the test section, and a constant speed motor may be used to drive the blower.

The 10,000 horsepower 1-by-3-foot supersonic wind tunnel utilizes four 2,500-horsepower synchronous motors driving individual compressors. The 50,000-horsepower supersonic wind tunnel now under construction will utilize two 25,000-horsepower wound-rotor induction motors driving an axial-flow compressor. However, wound-rotor motors were selected instead of synchronous motors, not for speed control, but rather from considerations of starting of large machines with high WR^2 load.

Variable Frequency Systems

In conjunction with the wind tunnels another major application of electric power is in the operation of model propellers on models undergoing tests. The rapid increase in the engine power in airplanes in recent years has made the effects of the propeller slipstream and propeller thrust so important an item in the characteristics and control of the airplane that it is no longer possible to obtain sufficiently useful test information on the model without duplicating the action of the propellers. In a full-scale wind tunnel it is usually convenient to use the actual service engine and propeller when testing airplanes. In models, however, it is necessary to provide some other source of propeller power. When a small-scale model is scaled down from the actual airplane, it generally is found that the space available for a motor to provide the motive power for the propeller on the model is much too small for any standard motor. Also, the rotational speed required for the propeller is very high. In scaling down the propeller operation from a full-scale airplane to a small model, it is necessary that the tip speed of the propeller on the model be the same as the tip speed on the full-scale propeller. This necessitates that the propeller speed be increased in proportion to the decrease in its diameter. Thus, if a full-scale propeller operates at 1,500 rpm, the propeller on a 1/10-scale model would have to turn at 15,000 rpm.

To meet these difficult conditions, the most practical solution has been the use of 3-phase squirrel cage induction motors, since this type of motor is simplest in construction and smallest in size for its horsepower rating. Even then, in order to get the power within the available space, all possible means must be taken to keep size at a minimum. Consequently, the motors are water cooled and are operated at temperatures which would not be tolerated in ordinary motors, but which give a reasonable length of life for such test purposes. A few small motors have even used carbon dioxide from high pressure cylinders as a refrigerant for cooling. As examples of what can be obtained in motors of this type, motors rated 22 horsepower at 12,000 rpm and measuring only $4^1/_2$ inches in diameter by 12 inches long and motors rated 100 horsepower at 10,000 rpm measuring 8 inches in diameter by $16^1/_8$ inches in length are typical of water cooled models. A motor 2 inches in diameter by $7^5/_8$ inches long is available rated 4 horsepower at 18,000 rpm with combined water and carbon dioxide cooling. Figure

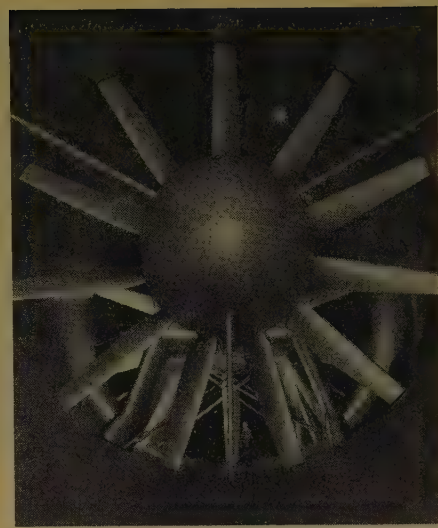

Figure 7. View looking upstream at fans of NACA 16-foot high speed wind tunnel at the Ames Aeronautical Laboratory—two fans, each direct connected to a 13,500-horsepower motor

Figure 8. Control motor – generator sets for NACA 16-foot wind tunnel drive system at the Ames Aeronautical Laboratory

Figure 9. Switchgear for control of the NACA 16-foot wind-tunnel drive at the Ames Aeronautical Laboratory

10 shows the 4-horsepower and 100-horsepower motors with a conventional 1-horsepower general purpose motor for comparison. These motors for models usually are designed to carry their rated load for approximately 20 minutes, which allows sufficient time for the average test. They have been built in ratings from 2 horsepower at 18,000 rpm to 1,000 horsepower at 2,400 rpm. The aeronautical laboratories continually are attempting to produce smaller and more powerful motors.

On these high-power high-speed motors, bearings are a major problem. Thus far, antifriction bearings have been used exclusively. A major addition to bearing troubles arises from the dissipation of the heat from the rotor of the motor. The stator is readily water-cooled, but there is no way of applying water to the rotor. Consequently, any heat generated in the rotor tends to travel along the shaft and heat the bearings. The point has not been reached where the bearings can be depended upon completely. However, it has been found necessary to provide the bearings with considerable internal looseness of the balls in the races in order to allow the inner race to expand without causing the bearing to become tight.

In nearly all tests using model propellers, it is essential to be able to vary the propeller speed over a wide range. Since the motors are of the squirrel cage induction type, the only means available for varying their speed is to change the frequency of the power supply from which they are operated. Thus, a variable frequency power supply is a necessary accessory for power-on tests. These variable frequency power supplies consist of either an alternator or a frequency changer driven at any speed over a wide speed range by a d-c motor supplied from a separate d-c generator with adjustable voltage control. In order to obtain the necessary high speeds with squirrel cage induction motors, it is necessary to operate them on high frequencies. Thus approximately 300 cycles per second are required for 18,000 rpm. There are seven separate variable frequency systems at the Ames laboratory of ratings from 100 kw at 400 cycles to 1,800 kw at 150 cycles. The motors usually are operated at a constant ratio of voltage to frequency because this results in a constant magnetic flux in the motor and gives approximately a constant value of current for a given torque throughout the speed range. However, different values of volts per cycle are required for different motors.

The distribution of the output of each variable frequency set to several points of utilization offers interesting problems. The ordinary a-c power distribution system is a simple parallel system—the voltage and frequency are constant, the circuit is energized at all times, and motors and equipment necessarily are provided with any features required for starting on a full voltage system. Any number of loads can be manipulated independently without appreciably affecting each other. On variable frequency systems, however, only a single load can be served at a given time from one variable frequency set since each load requires individual control of voltage and frequency. Controls must be provided at each point of utilization for varying the voltage and frequency, and interlocking must be provided to insure that only one station has control at any one time, all other loads being disconnected. Variable frequency motors normally are started on reduced frequency and voltage without auxiliary starting equipment. This necessitates interlocking equipment to prevent application of the power unless the voltage and frequency are at low values. These requirements result in rather involved systems when one variable frequency set must serve several different load stations, or when provisions are required for paralleling sets for added capacity. Distribution problems are complicated further by high voltage drop in cables because of high reactance at the high frequencies involved.

The same line of high speed motors and variable frequency supply systems is used to drive blowers, instead of propellers, to simulate the jet from a jet engine. The testing of jet propulsion combinations has introduced a new need for extremely high speed motors in high horsepower ratings.

Power Supply and Distribution System

Electric power is delivered to the laboratory at 110,000 volts from the Pacific Gas and Electric Company's large sub-

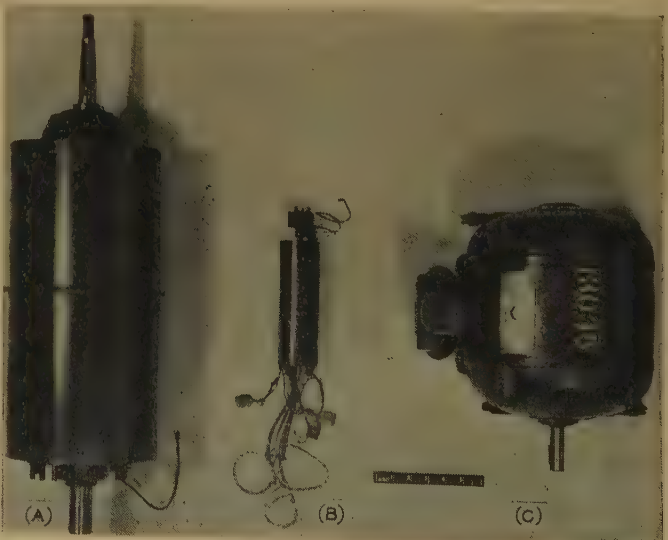

Figure 10. Two typical motors for powered aircraft models shown in comparison with a conventional general purpose motor

A. A 100-horsepower 10,000-rpm water cooled motor

B. A 4-horsepower 18,000-rpm water- and carbon-dioxide-cooled motor

C. A 1-horsepower 1,725-rpm conventional general purpose motor (shown for comparison)

Fault-Current Measuring Device

MARTIN J. LANTZ
ASSOCIATE AIEE

IN electric power system operation numerous problems are encountered in analyzing the results of short circuits and in the physical location of the trouble. Some devices now available for analyzing short circuits in a-c circuits are the oscillograph, the annunciator ammeter, and a magnetic link device requiring rectifiers. The oscillograph is the most accurate device but is expensive. The annunciator ammeter provides a limited amount of data; however, it offers a high electric burden to the circuit and does not have closely graduated steps of reading.

When a short circuit occurs on a power system, the current normally increases immediately in the faulted area. The device that records the phenomena must be very rapid and positive in action. It must not interfere with the electrical characteristics of protective apparatus in the same circuit, and it must be prepared to operate automatically, recording values that may be interpreted later. To be extensively applied, the device must be inexpensive, reasonably accurate, and reliable.

The fault-current measuring device described in this paper meets these requirements.

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Description of Device

The fault-current measuring device for use in a power circuit consists of one or more coils of insulated copper wire with magnetic links snugly fitted within them, and a standard high-speed voltage relay with low resistance and normally closed contacts shunting the coils (Figures 1–3). The use of a surge-crest ammeter and a demagnetizing coil is required to read the magnetization of the links and to demagnetize them after energization. The surge-crest ammeter is a special device used principally in lightning investigation for measuring the degree of magnetization of magnetic links. The magnetic link consists of three or four small magnetic laminations enclosed in a small tube. The device may be built with one or more coils to cover the desired current range. It is designed to operate in the secondary circuit of current transformers and is placed in series with the regular current-operated system protective relays or in the metering circuits. The burden of the device is low.

The high-speed by-pass relay is used with the device to minimize the effect of the initial d-c offset that may occur when the magnitude of an inductive alternating current is suddenly increased. In the field test application the by-pass relay was energized from the same direct voltage circuit that energized the tripping circuit of the oil circuit breaker. The by-pass relay which operates as a result of action by the system protective relays then opens its normally closed contacts across the coils one cycle or more prior to the zero point of the interrupted wave, allowing the current to be measured to flow through the coils and magnetize the links. The magnetic links are in the electric circuit when used with the by-pass relay only after the relay is energized and the by-pass contacts are open. Figure 4 is a typical schematic installation of the device. After operation of the device a target shows, the coils are again by-passed, and the by-pass relay locks out to prevent a second operation. A second device may be placed in service automatically if desired.

One device designed and tested covers a current range of approximately 4 to 60 amperes rms alternating current and consists of a 12- and a 56-turn coil in series. Assuming a 600/5 ratio current transformer, a primary current range of approximately 500 to 7,200 amperes is covered.

Theory

The magnetic link is placed in the coil and is magnetized by an alternating current flowing through the coil. This alternately magnetizes the link in one direction and then the other, the magnetism following the link's hysteresis loop. Each half cycle of symmetrical current demagnetizes the magnetism resulting from the previous half cycle and remagnetizes the link in the opposite direction. The electrical principle of the device is based upon the hysteresis loop of the small magnetic link and the interruption of the current at the zero point

station at Newark, Cal. A 50,000-kva and a 5,000-kva transformer bank reduce the voltage to 6,600 volts for distribution to the various loads which all are located relatively close to the substation. For the 50,000-horsepower supersonic tunnel under construction, an additional 62,500-kva transformer bank is being installed.

Owing to the large concentration of power, careful consideration has been required to keep the short-circuit capacity on the 6,600-volt feeders under 500,000 kva. To exceed the 500,000-kva limit would increase the cost of motor-control switchgear greatly as well as increase the probable extent of the damage in case of a fault. To keep within the 500,000-kva limit it was found necessary to avoid interconnecting the 6,600-volt busses of the proposed 62,500-kva bank and the existing 50,000-kva bank and install a small amount of reactance in the feeders to the 50,000-horsepower supersonic tunnel. The machines of the large wind tunnel drive systems make appreciable contributions to the short-circuit capacities on the 6,600-volt busses.

Power is distrubuted at 6,600 volts to the various buildings for small power and lighting. Y-connected 120/208-volt systems, on which 120-volt single phase loads may be distributed on all three phases, are used to advantage in the buildings, where the loads are a combination of lights and 3-phase motors. In a new hangar and shop building of much larger size than previous buildings, a 440-volt system was installed for building distribution and for motor loads with 440- to 120/208-volt dry-type transformers at load centers for lighting and convenience outlets.

Wind tunnels always operate intermittently, and much of the time at less than maximum rated power. Consequently, by staggering operation of several wind tunnels, it is possible to operate under a maximum kilowatt demand limit much less than the total kilowatt rating of the apparatus. In order to make most efficient use of power within the established maximum demand limit without unnecessarily delaying any wind tunnel, a totalizing and telemetering system is being installed which will indicate at all times at each wind tunnel control desk and at a central dispatcher's office the total kilowatt demand for the entire laboratory.

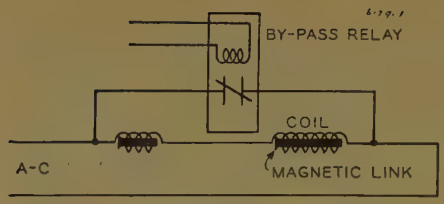


Figure 1. Schematic diagram of basic device for fault-current measurement

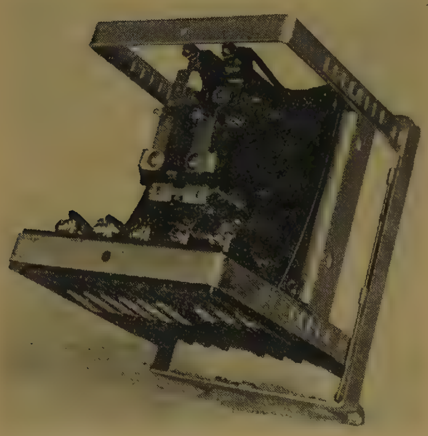


Figure 2. Rear view of preliminary sample of fault device showing relays

Device is mounted in cradle of standard single element protective relay

Figure 3. Front view of preliminary sample of fault device showing coils and links

(Figure 5). The residual magnetism retained by the magnetic link is proportional to the peak or crest magnitude of the last half cycle of the current, providing the last two half cycles are approximately equal in magnitude. This magnetism is then measured with a surge-crest ammeter. A calibration curve is obtained by plotting surge-crest ammeter readings against alternating current values. This provides a direct means for determining the last half cycle of alternating current through the coil by interrupting the current at zero and reading the degree of link magnetization by the surge-crest ammeter.

If the current is asymmetrical containing a d-c component, the magnetism of the link is affected and records a value which will not compare with a calculated or measured symmetrical value. The initial surge of direct current may place an amount of magnetism in the link that cannot be wiped out before the circuit is interrupted by the successive half cycles of opposite direction. The by-pass relay effectively eliminates the d-c component from the link as long as it is by-passed and then applies a reduced d-c component to the link, resulting in a more accurate determination of the magnitude of the last half cycle. By placing in series with the coils a 0.2- to 0.3-ohm resistance, the current is diverted to the by-pass relay contacts. This is a precaution against contact resistance and adds only a small burden to the circuit during the time the by-pass contacts are open. The illustrated replica of a test oscillogram demonstrates the action of this by-pass relay (Figure 6).

Use and Advantages

The basic device, which is considered to be the coil and link, will record automatically the magnitude of the last half cycle of an alternating current wave preceding its interruption at current zero by an oil circuit breaker. To obtain data of practical value on power systems with high-speed oil circuit breakers which may not allow time before opening for the d-c component effect to neutralize itself in the link, the by-pass relay is a necessary adjunct to the device. The data obtained by the device will aid greatly in determining the actual current values from short circuits on power systems.

Table I. Summary of Laboratory Tests on Fault-Current Measuring Device

Number of Turns in Coil	Number of Readings	Max Deviation, Per Cent	Average Deviation, Per Cent	Current Range, Amperes
45	6	−10.7	+2.1, −6.3	5.3–17
20	4	+ 8.7	+4.5, −4.2	9.4–38
12	12	+ 9.8	+3.3, −4.8	15.6–61
10	7	−10.6	+5.3, −7.8	18.8–77
5	6	+10.5	+8.7, −5.3	38.5–84

The following is a list of some of the advantages over devices which accomplish similar purposes, such as the automatic oscillograph and the annunciator ammeter:

1. Simplicity. There are no moving parts in the recording portion of the device, therefore no mechanical failures resulting from restrained movement can occur. The by-pass relay is considered as auxiliary to the basic device.

2. Instantaneous operation. There is no mechanical inertia in the recording portion of the device. It requires no mechanical action and is ready at all times for instant record.

3. Durable record. It is desirable to read the device soon after its operation but the reading may be deferred a number of hours.

4. It records a value proportional to the magnitude of alternating current without resorting to initial rectification.

5. Used in conjunction with a by-pass relay to record the last half cycle of a transient asymmetrical wave, the effect of the possible high initial d-c offset resulting from short circuits is materially reduced.

6. The device has a low electric burden and does not jeopardize the operation of protective devices.

7. The device may be automatically locked out of the circuit after operating.

8. The readings obtainable from the device are a continuous graduation of values obtained from a calibration curve allowing more accurate analysis to be made than is possible from a step-reading device such as the annunciator ammeter.

9. The device is inexpensive, making it

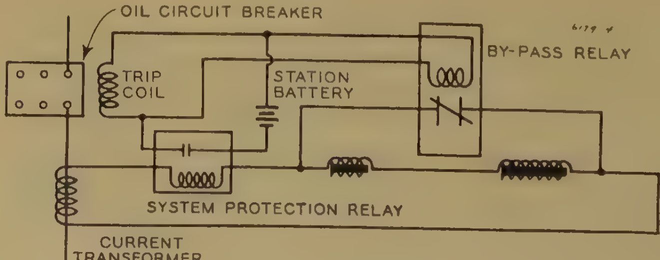


Figure 4. Schematic diagram of typical installation of fault-current device

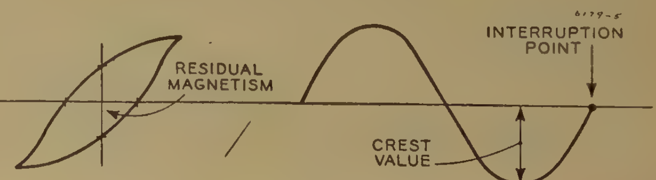


Figure 5. Hysteresis loop and sine wave

possible to distribute them extensively about the system to obtain operational records. As a fault-locating device and indicator of the type of fault, it can be applied to many lines where a more expensive device is not feasible.

Results of Tests

A series of tests using the magnetic link was made by the Bonneville Power Administration's electrical laboratory. These constituted both laboratory and field tests.

Laboratory Tests

The laboratory tests provided a-c and d-c magnetic link calibrations for various turn coils and tests to determine the effectiveness of the by-pass relay on asymmetrical waves. Data were secured for 5-, 10-, 12-, 20-, and 45-turn coils.

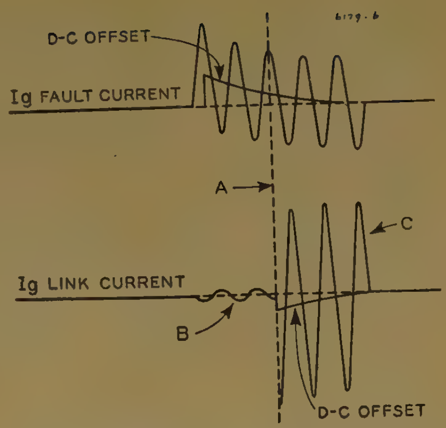


Figure 6. Replica of oscillogram showing results of use of by-pass relay across coil during system test

A—Current through coil when by-pass relay contacts are opened

B—Current through coil when by-pass relay contacts are closed

C—Last one-half cycle current measured by link

The a-c calibration data were obtained by adjusting the current with a variable resistive load. The energized circuit was closed through a circuit breaker into a resistance load, the circuit breaker opening at the zero point of current. Oscillographic records were obtained for all tests so that the results could be correlated. The a-c calibration points are indicated in Figure 5 by the circles. Peak amperes divided by the square root of two are plotted against surge-crest ammeter readings.

To determine the effect of the by-pass relay on offset currents, the resistance load was replaced with an inductive load. The current was passed through two coils in series, one of which was by-passed for a few cycles. The points in the squares, Figure 7, indicate the link readings containing the effect of d-c offset. The points in the triangles indicate the link readings in the by-passed coil. Shunting most of the d-c offset current around the link causes the link reading to move closer to the calibration curve.

Similar data are shown in Figure 8 for the 12- and 45-turn coils. Since the surge-crest ammeter readings are more accurate above 18 and below 70, these values were used as the limits of reliability.

Figure 7. Fault device laboratory test on 5-, 10-, and 20-turn coils

⊙—A-c calibration
△—Link reading short-circuited by relay
⊡—Link reading not short-circuited by relay

Curve	Link	Turns of Coil	Max Deviation, Per Cent	Avg Deviation, Per Cent	Current Range, Amp
A	8	20	8.7	4.5, −4.2	9.4 to 38.5
B	6	10	10.6	5.3, −7.8	18.8 to 77
C	4	5	10.5	8.7, −5.3	38.5 to 84.8

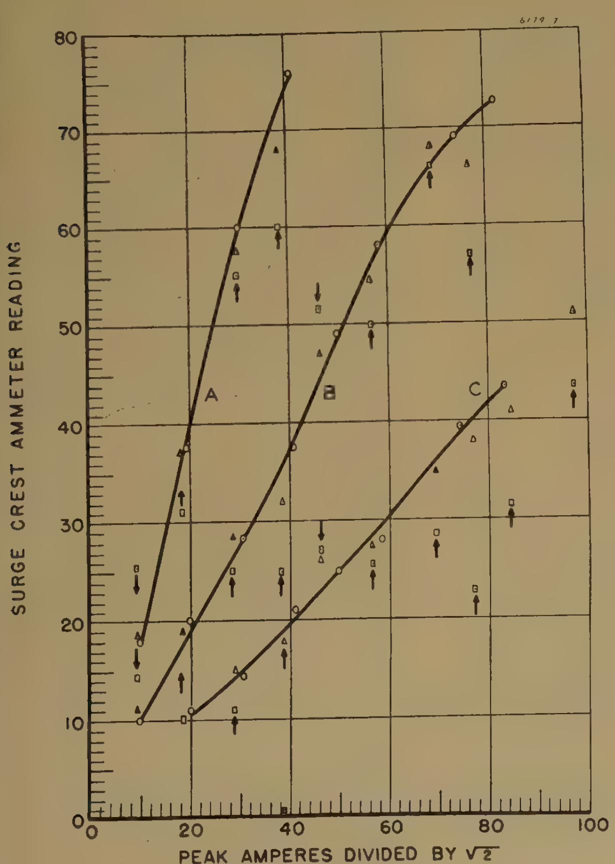


Figure 8. Fault device laboratory test on 12- and 45-turn coils

⊙—D-c calibration
△—Link reading short-circuited by relay
⊡—Link reading not short-circuited by relay

Curve	Link	Turns of Coil	Max Deviation, Per Cent	Avg Deviation, Per Cent	Current Range, Amp
A	8	45	10	2.1, −6.3	5.3 to 17.2
B	6	12	9.8	3.3, −4.8	5.6 to 61.1

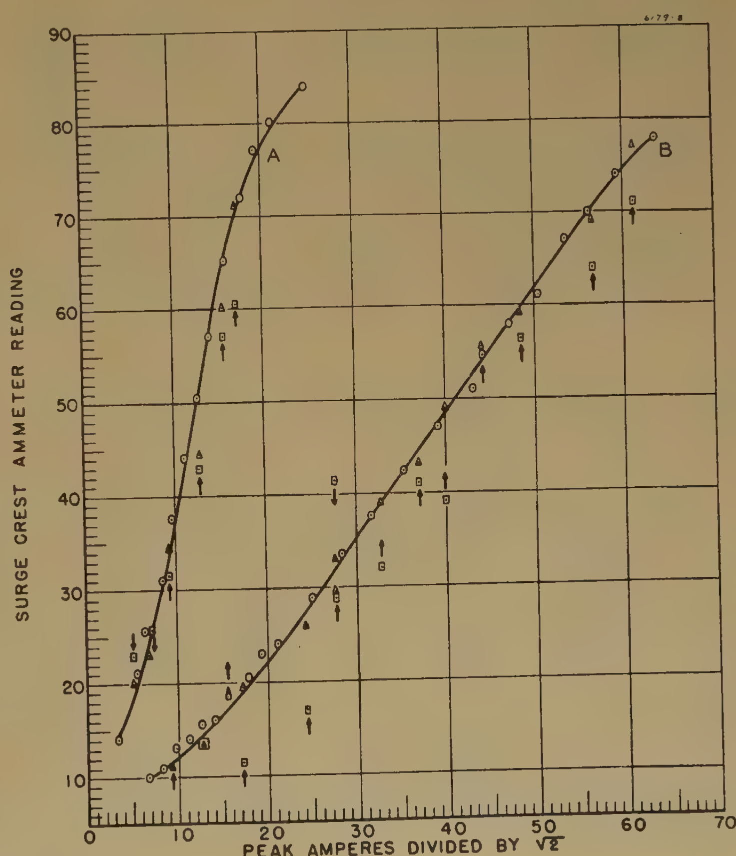
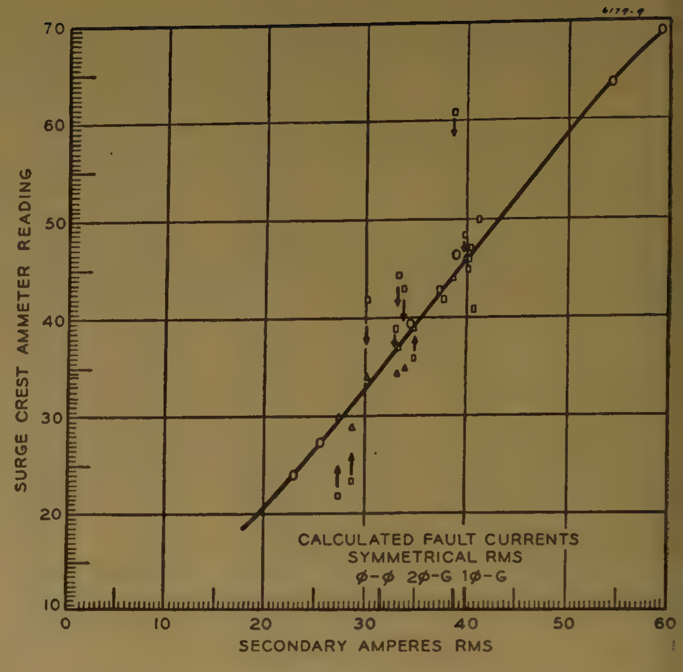


Figure 9. Fault device field test on 12-turn coil

⊙—A-c calibration data
⊡—Link reading without by-pass
△—Link reading with by-pass relay



The relationship between actual oscillographic measured current values and current values measured by means of the magnetic link are expressed in Table I as percentages of the link reading.

Theoretically the magnetic link should be magnetized in proportion to the peak value of current for a sinusoidal wave. Tests show the magnetic link d-c calibration curve to lie slightly higher than the corresponding a-c calibration curve which is to be expected after the first half cycle of magnetization.

All tests conducted should give link readings corresponding to the a-c calibration curve. However, the tests conducted in the laboratory corresponded more closely to the d-c calibration curves. This was probably the result of a higher percentage of harmonics introduced by the highly reactive test load used.

Tests in the field corresponded more closely to the a-c calibration curve as expected. The a-c calibration curve was used as the basis for determining the field test currents.

Field Tests

A series of staged system fault tests was made in May 1945, including single-phase-to-ground, two-phase-to-ground, and phase-to-phase faults. A fault-current measuring device with a by-pass relay was installed during the tests. The data resulting from these tests are shown in Figure 9. The test data were plotted on the a-c calibration curve for the 12-turn coil. All data were compared with oscillographic records taken during the test.

Table II. January 20, 1946, Midway Substation Single-Line-to-Ground Short-Circuit Tests

Results Obtained Using Fault-Current Measuring Device With By-Passed 12- and 45-Turn Coils

Test Number	Amperes Rms Last Half Cycle: Magnetic Link	Amperes Rms Last Half Cycle: Magnetic Oscillograph	Per Cent† Deviation
1	4,960	5,250	−5.9
2	4,770	4,770	0
2R	4,860	4,680	+3.7
*3	3,700	5,840	
4R	5,380	5,220	+3.0
5	1,295	1,290	+0.3
6	1,295	1,305	−0.8
7	2,305	2,030	+11.8
7R	2,400	2,170	+9.5
8	1,655	1,660	−0.3
8R	1,655	1,660	−0.3
9	1,645	1,845	−12.1
10	1,260	1,190	+5.6
		Average	−3.9, +4.8

* Failure to record correct value. Possible explanation in the text under field tests.

† $\frac{I_{\text{oscillograph}} - I_{\text{link}}}{I_{\text{link}}}$

The oscillograms of the tests show a higher degree of d-c offset present than normally would be expected for system faults. Several tests resulted in a rather large amount of d-c offset still present at the time of interruption.

A second field test was conducted in January 1946, during which thirteen single-phase-to-ground faults were made on the system. The results of this test are summarized in Table II. A maximum deviation from the oscillographic record was 12.1 per cent with an average deviation of minus 3.9 per cent and plus 4.8 per cent. One failure was recorded. A possible explanation for this failure was that a d-c transient caused by the collapsing flux of a saturated current transformer demagnetized the link causing it to read low. A d-c transient of opposite direction to the last half cycle shows on the oscillogram after current interruption.

Magnetic Link Data. Midway Test May 20, 1945

Type of Fault	Number of Readings	Max Deviation,* Per Cent	Avg Deviation,* Per Cent
1φ-G	5	−10.5	2.7, −3.7
2φ-G	5	− 2.3	0.9, −1.5
φ-φ	6	6.6	3.3, −4.2

* Compare to asymmetrical rms current values.

Conclusions

The magnetic link fault-current measuring device shows excellent correlation between its readings and corresponding values from oscillograms for both laboratory and field tests. The accuracy displayed on all the tests indicates a maximum deviation from the oscillographic record of approximately plus or minus 12 per cent and an average deviation of approximately plus or minus 5 per cent for the accepted design.

The tests conducted utilized a magnetic link and a surge-crest ammeter which were designed for a different purpose. Possibly a redesigned magnetic link and magnetism detecting device could be developed that would improve the operation of this device for the application to a-c measurements.

The device may be used extensively on a system because of its simplicity and low cost.

Since it is a current-operated device, it may be used on lines or feeders of any voltage using the proper current transformers.

There are several advantages of the device, one being improvement over the accuracy of the annunciator ammeter for use as a fault locator using the current-ratio method. It may be used as a check on relay target operation when additional coils are used in the separate phases to determine the type of fault and phases involved. It is believed the device has very good capabilities for use on power systems for the determination of the magnitude of fault currents. It provides a method for determining the interruptoin current of an oil circuit breaker during a fault. The latter may be of value in scheduling maintenance for oil circuit breakers. The device measures the magnitude of the last half cycle of current being interrupted. The difference between the magnitude of the last half cycle cur-

Frequency Performance of Thyratrons

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Synopsis: The frequency performance of small thyratrons is investigated by supplying anode voltage from a variable frequency electronic generator of 1,400 watts output capacity. At high audio frequencies there is a departure from the 60-cycle-per-second performance; the grid control characteristic resolves itself into two characteristics—a starting characteristic and an extinguishing characteristic. The starting characteristic is shown to be a function of grid–anode capacitance and grid resistance. The extinguishing characteristic is determined by deionization effects and so is a function of frequency, grid resistance, anode current, tube geometry, and gas pressure.

THERE are many applications for thyratrons at frequencies above the power frequency of 60 cycles per second. Inverters, servo controls, relaxation oscillators, radar modulators, and grid-controlled rectifiers for airborne equipment are examples of such applications. The frequency range is from a few hundred to 50,000 cycles per second. Although the circuits used in these applications are well understood, the exact nature of ionization and deionization effects on thyratron performance generally has not been known. This paper reports an investigation of such effects as affected by anode-supply frequency. A sinusoidal wave shape was used for the various tests because it frequently is used in actual circuits and because it is easy to generate. Wave shape affects thyratron performance, but performance with other wave shapes can be extrapolated from the results obtained with sine waves.

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In the tests the sinusoidal voltage is applied to the anode circuit of the tube under test; hence the voltage source must have considerable power capability. A high-power electronic audio-frequency generator capable of supplying 1,400 watts of power at 115 volts in the frequency range of 300 to 3,500 cycles per second was used in this investigation.[1] By means of an external signal generator, the equipment also could be used as a power amplifier to produce reduced power at frequencies up to 20,000 cycles per second. The output voltage of the generator is maintained at a preset level independent of load by means of a regulating circuit having a time constant of about one second.

Figure 1 shows a block diagram of the test setup. T is a stepup transformer to supply the anode voltage for the tube under test. R_L is a load resistor to limit the average anode current $\bar{I}_b$. Filament power for the tube under test comes from the 60-cycle power line. Peak anode voltage is measured with a diode type peak voltmeter. Wave forms of both current and voltage are monitored on a cathode-ray oscilloscope. Grid voltage is supplied by a 90-volt battery.

One type of thyratron, 3D22, was selected to be run through all the tests in which conditions would be subject to control by the user. The type 3D22 is a xenon-filled shield-grid thyratron rated for a peak forward voltage of 650 volts, a peak inverse voltage of 1,300 volts, and an average current of 0.75 ampere. Its control ratio is 150.

However, other tube types were put through some of the tests. The results reveal the effects of tube geometry and aid the user in selecting the proper tube type for a given application.

An additional test shows the effect of change in gas pressure, a characteristic which is subject to control by the user only in the case of mercury thyratrons.

Thyratron Control Phenomena

Transition to On-Off Control

At a low frequency (60 cycles per second), the average current through a thyratron can be controlled by the grid and can be calculated from the following equation,[2] which neglects tube drop:

$$\bar{I}_b = \frac{\hat{E}_{bb}}{2\pi R_L}(1+\cos\theta)$$

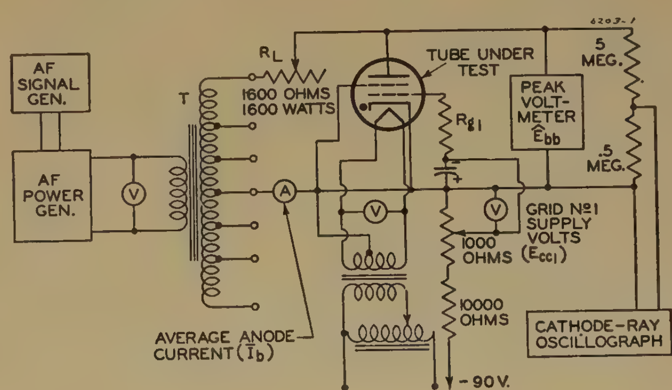
$\hat{E}_{bb}$ = peak anode supply voltage (sinusoidal)
R_L = load resistance
θ = phase angle at which conduction begins

When the grid voltage is zero in a negative grid thyratron, θ is zero, conduction starts at the beginning of the anode voltage cycle, and maximum average current flows. With increasing negative grid voltage, θ approaches 90 degrees and the current is reduced to one-half the maximum average value. With only direct voltage on the grid, 0 to 90 degrees is the maximum range of variability since making the grid still more negative cuts off the current completely.

When the anode supply frequency is increased sufficiently, this mechanism breaks down. The grid no longer is able to vary the anode current; it becomes simply an on-off control, the current being established only by the external anode circuit. Figure 2 shows this transition in the case of the type 3D22 thyratron. The explanation of this form of operation is that ions remaining from the previous conduction cycle interfere with the grid control. At low frequency the interference is small and the extinction current is

Figure 1. Test setup for audio frequency testing of thyratrons

$\bar{I}$ = average current
$\hat{E}$ = peak voltage
Rating of transformer T:
Primary = 115 volts
Secondaries = 3,500/1,750 volts, 2,000/1,000 volts, and 1,000/500 volts
Volt-amperes = 2,120/2,750
Frequency = 0.3 to 3.5 kc



rent and the current at the time the contacts part, which is defined as the interrupted current, is usually negligible. This may not be true in the case of very fast circuit breakers opening before the d-c component has decayed to a noncritical point, for restrikes, nor for low-voltage circuit breakers where the arc voltage is sufficient to choke down the last half cycle of current. The instances in which the device will be subjected to these conditions probably will be infrequent.

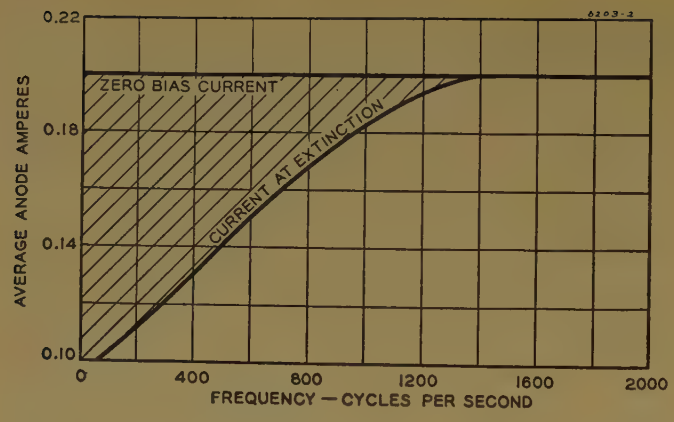


Figure 2. Extinguishing current versus frequency for type 3D22 thyratron

$R_{g1}=0.1$ megohm
$E_{cc2}=0$
$R_{g2}=0$
$\hat{E}_{bb}=650$ volts

therefore nearly half the maximum current. As the frequency is increased, this interference increases, the extinction current increases until it reaches the maximum average value, and we have solely on-off control.

However, if one desires to obtain a variable anode current at frequencies beyond 1,500 cycles per second for the type 3D22 thyratron one can use the d-c bias-phase-control method.[3] In this case the grid voltage consists of a combination of direct voltage plus a 90 degrees lagging alternating voltage. This method is successful in controlling the average anode current at higher anode supply frequencies whereas the direct voltage method alone would fail. The explanation is that the alternating grid voltage sweeps out ions while swinging negative. Figure 3 gives the conditions under which this control may be applied to the type 3D22 thyratron at a frequency of 2,000 cycles per second where d-c grid voltage alone would not permit variable control.

Two Control Characteristics—Starting, Extinguishing

It became apparent as these tests progressed that there is not one but two control characteristics. One characteristic describes the relation between anode voltage and grid voltage on starting.

Table I. Capacitance Measurements of the Type 3D22 Thyratron

Tube Component	Capacitance, μμf
Socket	0.229
Grid–anode (inside envelope)	0.094
Effective base pin (tube in socket)	0.126
Total	0.449

This is the initial conduction of current and is not to be confused with the every cycle starting which accompanies an a-c anode supply. The other describes this relation for extinction or cutoff of the discharge. By cutoff is meant failure to reignite on successive cycles. Figure 4 shows how the extinguishing characteristic has shifted considerably in the negative direction at 2,500 cycles per second. This shift is caused by a lack of complete deionization. At a low frequency (60 cycles per second) the extinguishing characteristic is coincident with the starting characteristic.

To the author's knowledge the first investigator to indicate that there are two control characteristics was W. G. Shepherd of the Bell Laboratories. In 1943 Shepherd published a paper describing work on deionization effects in thyratrons.[4] He differentiated between a static and dynamic characteristic. The work wa[s] done with alternating voltage on the grid and the dynamic characteristic was i[n] terms of peak a-c grid swing necessary fo[r] stable operation.

However, it is believed that the concept of extinction control characteristic as herein presented is closer to the accepted definition of control characteristic.[5]

The Extinguishing Control Characteristic

As would be expected the extinguishing characteristic is a function of frequency. Figure 5 shows how this characteristic shifts with frequency for the type 3D22 thyratron. At 3,500 cycles per second the grid cutoff voltage is nearly ten times the 60-cycle value. The shift is increasing at a rate faster than proportional to frequency. Nevertheless, the grid is still potent as a control electrode, and in none of these tests did the grid lose control completely. If the grid voltage is made sufficiently negative it will sweep out the ions at a rate such that it can prevent reignition on the next positive cycle. The limit is reached, however, when the sum of the negative grid voltage plus the peak anode voltage exceeds the cold breakdown value between grid and anode; in other words Paschen's law is operating.

The Starting Control Characteristic

The starting control characteristic for the type 3D22 thyratron was found to be constant for frequencies from 60 to 3,500 cycles per second. However, this condition is a special case; usually the starting control characteristic is a function of fre-

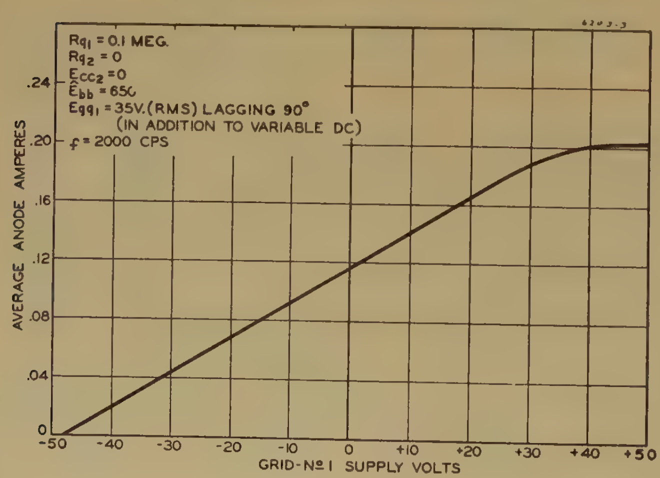


Figure 3. Thyratron control of type 3D22 combined direct voltage and 90-degree lagging alternating voltage

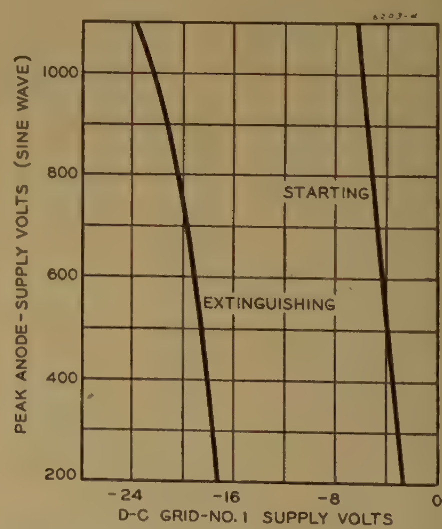


Figure 4. Control characteristics of type 3D22 thyratron starting and extinguishing

$R_{g1}=0.1$ megohm $R_{g2}=0$
$I_b=0.2$ ampere at $E_{cc1}=0$ $E_{cc2}=0$
Frequency = 2,500 cycles per second

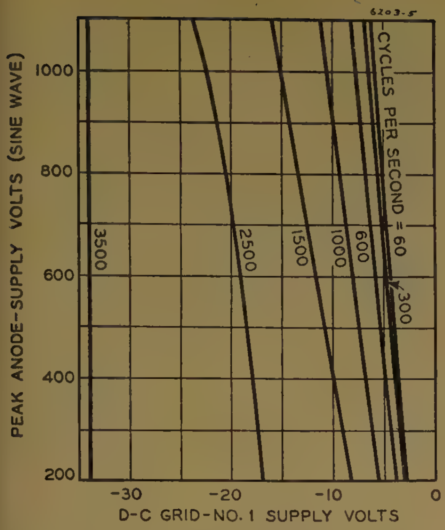


Figure 5. Typical extinguishing control characteristic of type 3D22 thyratron

$R_{g1}=0.1$ megohm $\quad R_{g2}=0$

$\bar{I}_b=0.2$ ampere at $E_{cc1}=0 \quad E_{cc2}=0$

quency. In order to illustrate the general case consider the type 2050 thyratron. The type 2050 is a xenon-filled shield-grid thyratron similar to the type 3D22 but smaller in size. The ratings are a peak forward voltage of 650 volts, a peak inverse voltage of 1,300 volts, and an average current of 0.1 ampere. Its control ratio is 250. Figure 6 gives the starting control characteristic as a function of frequency for this type. One may observe that, like the extinguishing characteristic, the shift is toward the negative grid voltage but that the shift is not as great. This shift is not caused by ionization but by coupling to the grid through the anode–grid capacitance shown in Figure 7.

The anode–grid capacitance C_{pg} is small and therefore its reactance is large in comparison with the grid resistance. Hence the induced grid voltage is nearly 90 degrees ahead of the anode voltage. This induced voltage becomes tangent to the critical grid voltage at point x. As the frequency of the anode supply is increased the induced voltage increases, and thus conduction begins at a more negative d-c grid voltage.

Control Versus Grid Resistance

The tests presented thus far all were made at one value of grid resistance, namely 0.1 megohm. Figure 8 shows the effect of varying this grid resistance at a fixed frequency of 1,000 cycles per second. Both the starting and extinguishing grid voltages become more negative as the grid resistance is increased.

It is now clear why no shift in starting characteristic was observed in the case of the type 3D22 thyratron. The grid re-

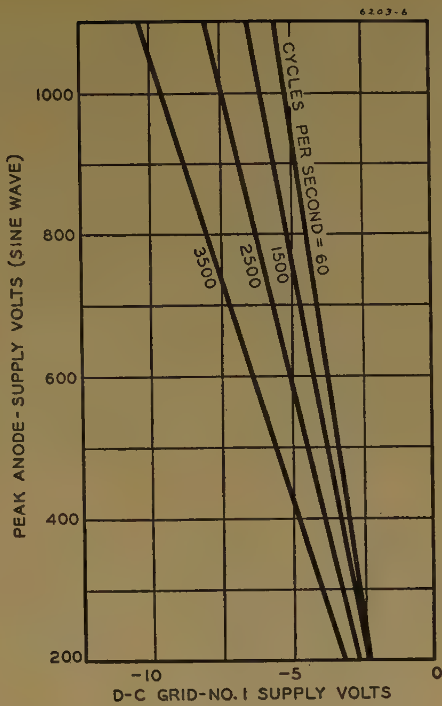


Figure 6. Typical starting control characteristic of type 2050 thyratron

$R_{g1}=0.1$ megohm $\quad R_{g2}=0 \quad E_{cc2}=0$

sistance chosen (0.1 megohm) was not sufficiently large.

The explanation of the starting shift has been given. To check the data in Figure 8, capacitance measurements were made on the type 3D22 thyratron and its socket with the results as given in Table I. It is interesting to note that only one-fourth of the capacitance is contributed by the tube electrodes. Tube capacitance as published by tube manufacturers includes neither the interbasepin capacitance nor the socket capacitance.

The capacitive reactance of 0.449 micromicrofared at 1,000 cycles per second is 354 megohms. The peak induced grid voltage for the conditions of Figure 8 at five-megohm grid resistance is 9.18 volts The induced grid voltage does not cut the

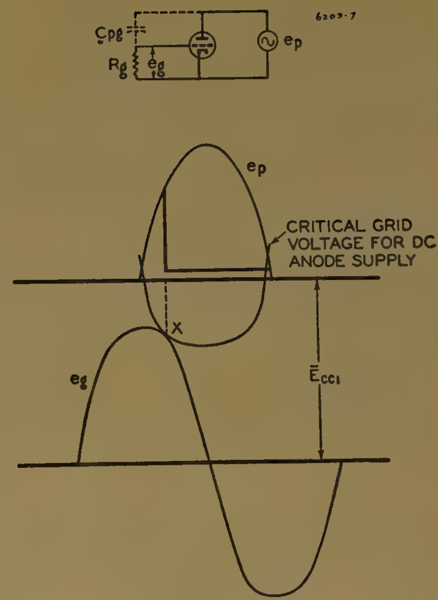


Figure 7. Anode–grid coupling in thyratron

critical grid voltage at the peak of the induced voltage, but it is close to this value. Reading the data from the curve of Figure 8, we find that the shift in control grid bias from 60 to 1,000 cycles per second is from minus 4.5 to minus 16.5 or 12 volts. The remainder of the shift is caused by direct current resulting from leakage or grid emission.

The shift in extinguishing characteristic increases with grid resistance because of the effects just described plus deionization effects. Increasing the grid resistance decreases the effectiveness of the grid as a deionizing agent. The grid is able to sweep out fewer ions between conduction periods.

Control Versus Anode Current

It is a common conception that the control characteristic is independent of current. At high audio frequencies this is not true. The cutoff bias is a linear function of the average anode current as Figure 9 shows.

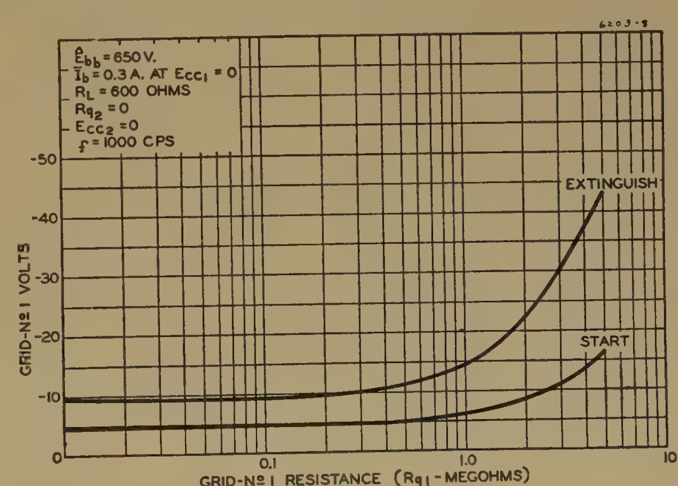


Figure 8. Critical grid bias versus grid resistance of type 3D22 thyratron

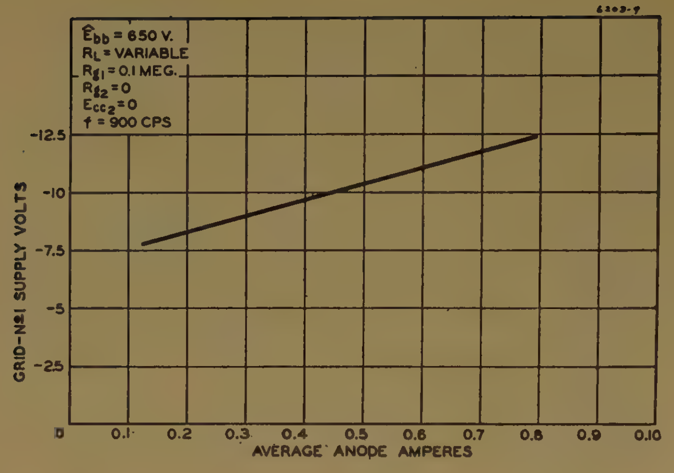


Figure 9. Extinguishing grid voltage versus average anode current for type 3D22 thyratron

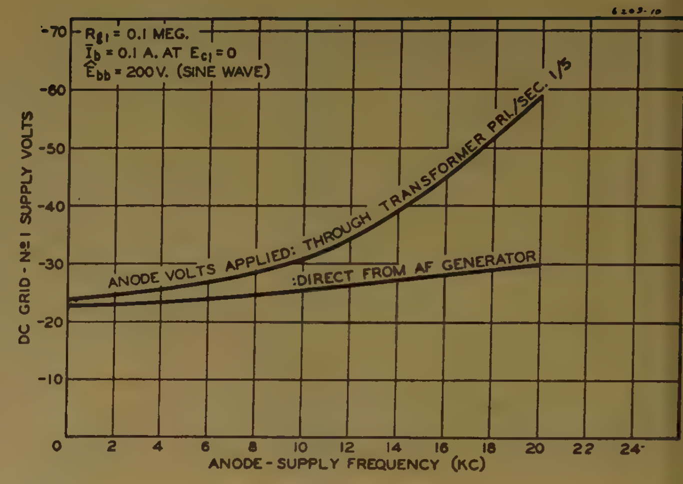


Figure 10. Effect of inductance in anode supply on extinction for type 884 thyratron

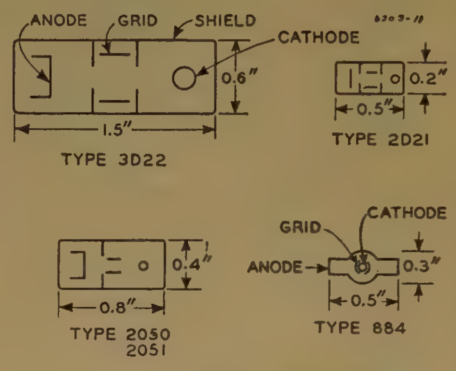


Figure 11. Thyratron constructions

Such a linear relationship would be expected if the following train of events are considered. The number of positive ions in the discharge is proportional to the current that has passed. Between conduction cycles the quantity of ions decays exponentially at a fixed rate, so that at a fixed time after current ceases in the external circuit the number of ions is proportional to the current that flowed. The negative grid, being the most effective deionization agent, sweeps the ions out at a rate approximately proportional to the negative applied voltage. At the beginning of the next cycle, if the grid has reduced the ions below a critical number, the anode rising in voltage will not be able to initiate a discharge.

Because of these relationships the anode current must be specified when the control characteristic at high frequency is given.

Effect of Inductance

When there is inductance in the load, the current persists after the anode voltage has dropped to zero. The solution[8] for a pure inductive load is

$$i_b = \frac{\hat{E}}{\omega L}(\cos\theta - \cos\omega t)$$

The generator produces a sinusoidal voltage of angular frequency ω and peak value $\hat{E}$. θ is the angle of firing as set up by E_{cc}, the d-c grid supply.

This persistence of current beyond 180 degrees results in a larger number of ions in the tube at the beginning of the next conduction cycle and hence requires a larger negative grid bias to sweep these ions out and to allow the grid to regain control. The effect may be seen in Figure 10 where a more negative curve was obtained when a transformer was inserted between the audio generator and the tube under test.

In taking these data the transformer ratio, resistive load, and voltage were selected to minimize the effect of transformer inductance.

Comparative Data on Different Thyratrons

Types 2051, 2D21, and 884 also were tested in this investigation in addition to types 3D22 and 2050. Ratings of these types appear in Table II. Figure 11 illustrates the construction of each one of these types. The cross sections are shown accurately to scale; the dimensions given are inside values. Note the similarity among the three types 3D22, 2050, and 2D21. Each of these types is a xenon-filled shield-grid thyratron. Types 2051 and 884 are argon-filled.

Starting Control Characteristic

In Figure 12 the starting control characteristics are compared for the four high-control-types. Types 2050 and 2051 have a high anode–grid capacitance and therefore exhibit a large shift. The 2D21 thyratron has the smallest grid–anode capacitance and hence the smallest shift. The frequency has to be increased to 15,000 cycles per second before a shift is observed in the 2D21.

Figure 14 gives the starting characteristic for type 884. This thyratron has a large grid–anode capacitance but its control ratio is so low that the shift is not excessive even at high audio frequencies.

Extinguishing Characteristic

The extinguishing control characteristic for the type 3D22 thyratron is shown in Figure 5. In Figure 13 the extinguishing characteristics are compared for the four high-control-ratio types. The shift in control is roughly proportional to the cross-sectional area of the type. Table III gives another indication that the phenomena are effects of deionization.

Although the type 2051 thyratron is a

Table II. Comparative Ratings of Thyratron Tubes

Type	Peak Forward Voltage, Volts	Peak Inverse Voltage, Volts	Average Current, Amperes	Anode–Grid Capacitance (Including Socket), $\mu\mu f$	Control Ratio
3D22	650	1,300	0.75	0.45	150
2050	650	1,300	0.100	0.60	250
2051	350	700	0.075	0.60	250
2D21	650	1,300	0.100	0.25	250
884	350	350	0.075	6.4	10

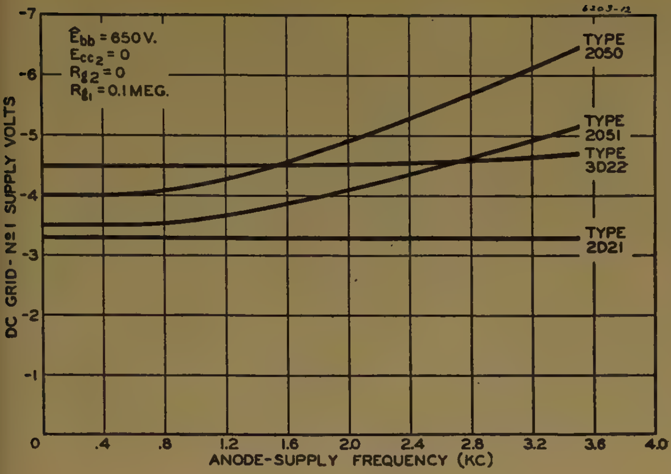


Figure 12. Comparison of starting grid voltages for types 2D21, 3D22, 2050, and 2051 thyratrons

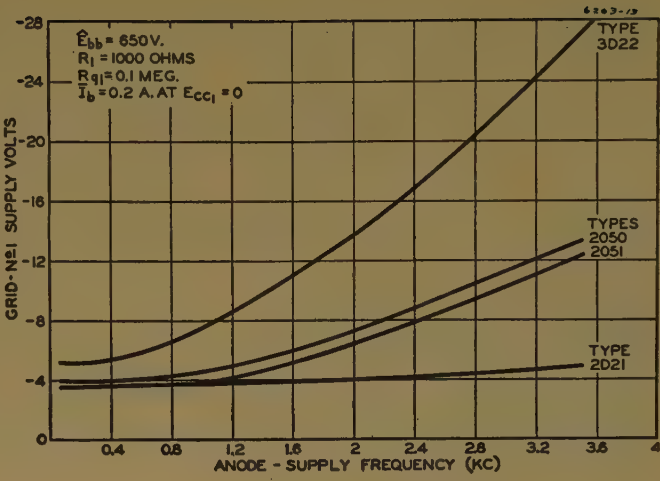


Figure 13. Comparison of extinguishing grid voltages for types 2D21, 3D22, 2050, and 2051 thyratrons

renewal type, not to be specified for new equipment, it was tested because it was expected that the use of the lighter gas argon in place of xenon would lower its deionization time. However, the gas pressure is nearly three times greater in the latter tube and therefore counteracts the difference between gases.

Referring to Figure 14 for the type 884, it may be seen that the starting and extinguishing characteristics are identical. One must conclude that the effect of deionization on the control characteristic is negligible even up to 20,000 cycles per second. The whole shift is caused by grid-anode coupling; consequently, the type 884 is the most desirable tube for high frequency work.

Effect of Pressure

Starting Control Characteristics

A number of type 2050 tubes were made with pressures ranging from 25 to 250 microns of xenon. The starting characteristics of these tubes were measured, but there is no definite shift with pressure. Had the pressure range been extended down to about five microns a shift probably would have been observed since this is the pressure range in which mercury thyratrons exhibit a large shift in control characteristic.[6] The explanation for this shift in starting control characteristic of a thyratron with pressure will be considered in order to distinguish it from the other shift phenomena observed in this paper. It is the same as the explanation for Paschen's breakdown curve of a gas diode. As the pressure is increased each electron makes more collisions and produces more ions in crossing the tube. Thus the critical anode current for the formation of a discharge with the given anode voltage is reached at a lower initial electron current or at a greater negative grid voltage.

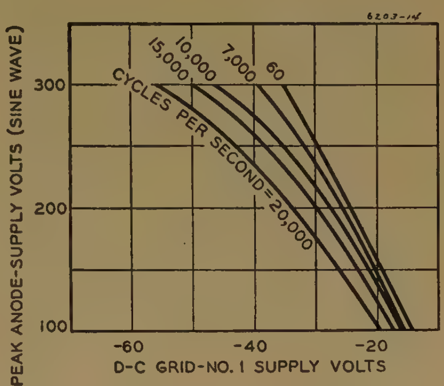


Figure 14. Starting and extinguishing characteristic of type 884 thyratron

$R_{g1}=0.1$ megohm $\quad \bar{I}_b=0.1$ ampere at $E_{cc1}=0$

Extinguishing Control Characteristic

It is known that deionization time is a function of pressure[7] so that, if the shift in extinguishing control characteristic is a deionization effect, the shift should be a function of pressure. The results of the test for extinguishing characteristic are elegant as Figure 15 will attest. The curves fall in order toward the negative grid voltage as pressure is increased.

Conclusions

The following conclusions can be drawn about the performance of thyratrons at high audio frequencies:

1. The grid loses control over the average anode current but is still potent enough to shut off the discharge.

2. There are two control characteristics instead of one—a starting characteristic and an extinguishing characteristic.

3. The starting characteristic shifts negatively with increasing frequency as a result of coupling through grid–anode capacitance.

4. The extinguishing characteristic shifts negatively with increasing frequency because of coupling through grid–anode capacitance as well as because of deionization effects.

5. The extinguishing characteristic shifts negatively with increasing grid resistance.

6. The extinguishing characteristic shifts negatively with increasing anode current.

7. A comparison of tubes of similar construction indicates that the shift of extinguishing characteristic with frequency is roughly proportional to the cross-sectional areas of the tube cage.

8. The starting characteristic may shift

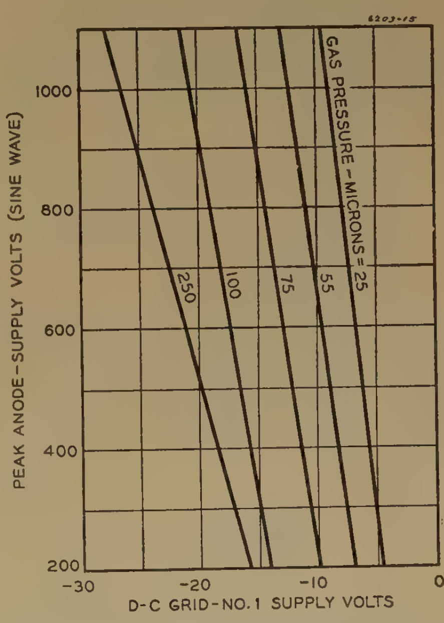


Figure 15. Extinguishing characteristic versus pressure for type 2050 thyratron

$R_{g1}=0.1$ megohm $\quad R_{g2}=0$
$\bar{I}_b=0.200$ ampere at $E_{cc1}=0 \quad E_{cc2}=0$
Frequency = 3,500 cycles per second

Low Reactance Flexible Cable for Induction Heating

MYRON ZUCKER
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THERE has been a demand for a flexible cable to carry single phase current for induction heating at frequencies up to 10 kc. Tests on an interleaved type of cable, used widely in portable spot welding, indicate its usefulness in carrying thousands of amperes at potentials under 1,000 volts at the higher frequencies.

The flexibility of the cable has been sought for two purposes in induction heating:

1. For convenience of installation in the generator-transformer-coil circuit and for occasional shifting of the machinery.

2. For ability to carry currents (generally at low voltages) to coils that may be frequently or even continuously in motion.

In some cases, maneuverability is so important that it is advisable to use water cooled conductors in order to minimize cable size. Especially in such cases, it is logical to consider an adaptation of an interleaved type cable now widely used to carry heavy currents to portable spot welders. The basic problems of low reactance, high flexibility, and water tightness have been solved in this cable. The principal new requirement for induction heating is higher electric insulation. This has been attained for several hundred volts, and the cable is being used with and without water cooling.

The purpose of this paper is not to describe the cable (which has been covered elsewhere from the welding viewpoint[1,2]), but to estimate and give test values for some physical and electrical characteristics pertaining to its use at higher frequencies. The material offered here is not exhaustive, but introduces the characteristics of this type of cable which appears to have advantages in economy, low reactance and resistance, and flexibility compared with coaxial cable. Interleaved cable should not be used where its external field would be troublesome, unless it is shielded. Coaxial cable, described in many articles where studies of coaxial cables were made for other purposes,[3,4,5] may be preferable in such cases.

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This presentation is limited to interleaved type cable. Physical dimensions are covered. They affect flexibility (cables of larger diameter being stiffer) and must be known for calculation of inductive reactance. The latter is covered in some detail since it is the major element of impedance of the cable at the frequencies involved. Some data are given on resistance, capacitance, and leakage losses. Cable ratings are not considered here since further information is needed on increased resistance losses at high frequencies in order to extend welding application data.[2] Also, more data are required on heating time constants at low water rates to evaluate heating and cooling of the cable under intermittent operations.[6]

Description of Cable

The essence of this cable design is that each conductor is broken up into a number of "ropes" (see Figure 1) placed around a central nonconducting core in a manner similar to the outer groups of the ordinary concentric-lay stranded electric cable. Ropes of alternate polarity lie next to each other except for a thickness of insulation between them. Since water cooling is desired each conductor is bare, and the insulation takes the form of a tube of irregular cross section that is formed in place inside half of the conductors and outside of the other half. The whole mass is held in a

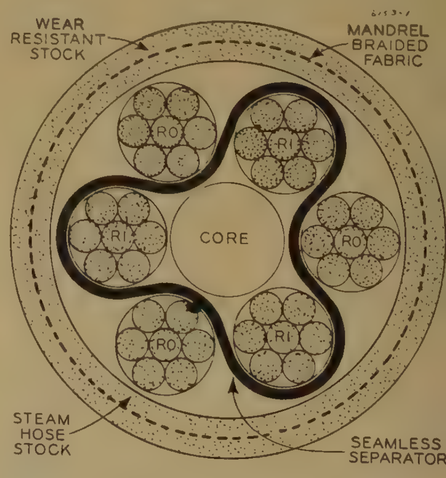

Figure 1. Cross section of interleaved cable for six ropes, three per electric side

negatively with increasing gas pressure because of increased ionization efficiency.

9. The extinguishing characteristic will shift negatively with increasing gas pressure because of increased ionization efficiency as well as for reasons of longer deionization time.

Appendix. Nomenclature

$\bar{I}$ = average current
$\hat{E}$ = peak voltage
E_{cc1} = d-c supply voltage for number 1 grid
E_{cc2} = d-c supply voltage for number 2 grid
R_{g1} = resistance in series with number 1 grid
R_{g2} = resistance in series with number 2 grid
E_{c1} = voltage at number 1 grid
E_{c2} = voltage at number 2 grid
E_{gg1} = a-c (rms) supply voltage for number 1 grid
$\hat{E}_{bb}$ = peak supply voltage for anode
$\bar{I}_b$ = average anode current
f = frequency of anode supply

Note: All voltages are referred to cathode.

Table III. Extinguishing Control Characteristics of Thyratron Tubes

Type	Cross Section, Square Inches	Shift in D-C Supply Voltage for No. 1 Grid E_{cc1}, Volts
3D22	0.818	22
2050	0.284	9
2D21	0.118	2

60–3,500 cycles per second; $\hat{E}_{bb}$ = 600 volts.

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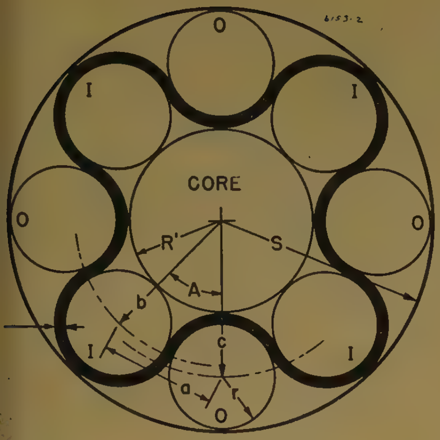

Figure 2. Symbols for analysis of cable section

regular pattern by means of an outer covering.

Some of the computations that follow are based on this arrangement. Then simpler formulas are derived from an approximate pattern in which all the conductors are on the circumference of one circle with radial insulating partitions. It is evident that the more ropes that are provided, the lower will be the reactance and the skin effect, which becomes important at higher frequencies. More ropes entail a larger core.

The core is a nonconductor which has been adopted for mechanical reasons even though the use of copper in the core would result in a smaller over-all cable diameter. The question is how great a price in terms of bulk must be paid for the

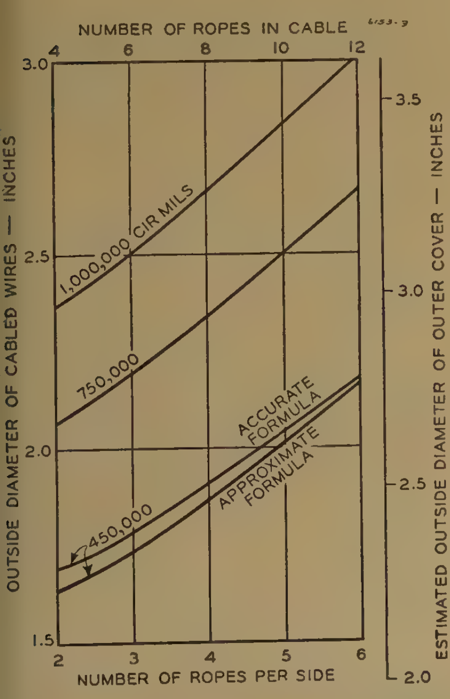

Figure 3. Calculated dimensions versus number of ropes in three sizes of interleaved cable

lower impedance obtained by subdividing the copper.

Cable Dimensions

The size of the cable is important not only because it determines flexibility, but also because the electrical characteristics depend on the size and the distance between conductors.

The geometry is somewhat complicated by the fact that the conductors are on the circumferences of two circles. The ropes O outside the insulating separator are on one circle; the ropes I that are inside are on another circle (see Figure 2).

Let

r = radius of each conducting rope

R' = radius of insulating core

t = thickness of insulating separator

Then let

a = straight line distance between centers of adjacent ropes $=2r+t$

b = radius of circle upon which I ropes center $=R'+r$

c = radius of circle upon which O ropes center $=R'+r+t$

To solve the triangle abc, let

$$S=\frac{a+b+c}{2}$$

Then

$$S=\frac{2r+t+R'+r+R'+r+t}{2}=R'+2r+t$$

which, by coincidence, equals the radius of a circle around the cable.

If n equals the total number of ropes, then the angle between the centers of adjacent ropes is $A=360°/n$.

Since

$$\sin\frac{A}{2}=\sqrt{\frac{(S-b)(S-c)}{bc}}$$

we obtain, by substituting for A, S, b, and c from the foregoing expressions,

$$\sin^2\frac{A}{2}=(\sin^2)\frac{180°}{n}=(S-b)(S-c)/bc$$

$$=\frac{(r+t)r}{(R'+r+t)(R'+r)}$$

This may be solved for the diameter of the core:

$$2R'=-(2r+t)\pm\sqrt{(2r+t)^2-4r(r+t)\left(1-\frac{1}{\sin^2\frac{180°}{n}}\right)} \qquad (1)$$

Then the diameter D of the composite, which is also the inside diameter of the outer covering (neglecting clearances required for assembly), may be expressed as follows in terms of r, t, and n:

$$D=2S=2R'+4r+2t$$
$$=2r+t+\sqrt{(2r+t)^2-4r(r+t)\left(1-\frac{1}{\sin^2\frac{180°}{n}}\right)} \qquad (2)$$

The values of D for several sizes of cable (with 4 to 12 ropes) are shown in Figure 3.

A simpler approach is to assume that the centers of the ropes are on the circumference of one circle of radius R. The circumference of this circle is

$$2\pi R=n(2r+t)$$

whence

$$R=n\frac{(2r+t)}{2\pi}^{*} \qquad (3)$$

The outer diameter of the array is taken as the diameter of this circle plus the diameter of a rope and one thickness of the separator. Thus

$$D=2R+2r+t$$
$$=n\frac{(2r+t)}{\pi}+2r+t \qquad (4)$$

The closeness of this approximation when insulation thickness is small compared with the size of conductors is shown in Figure 3. These curves are based on theoretical values of r. In actual cases sizes may vary considerably because of different stranding and cabling practices. The curves are within reason, however, and illustrate the effect of varying the amount and subdivision of copper area in this design.

Electrical Characteristics

Inductive Reactance

The induced voltage in each rope is the line-to-neutral reactive drop. Using the method of counting flux linkages up to some distance u outside of the cable, the voltage induced in rope 1 per centimeter length is found to be

$$E_{L-N}=\omega i 10^{-8}\times\left[\frac{1}{20}+0.2\ln\left(\frac{u}{r}\cdot\frac{u}{a_3}\cdot\frac{u}{a_5}\cdots\frac{a_2}{u}\cdot\frac{a_4}{u}\cdot\frac{a_6}{u}\cdots\right)\right]$$

where a_n is the distance between the center of rope 1 and the center of the nth rope and i is the current per rope.

* More accurately,

$$R=\sqrt[n-1]{\frac{a}{n}}=\sqrt[n-1]{\frac{2r+t}{n}}$$

Through the following eight simple transformations, we obtain the value of X_{LL}:

1. Change to common units of volts per foot (2.54×12 multiplier).
2. Line-to-line voltage $E_{LL}=2E_{LN}$.
3. Total current $I=ni/2$.
4. Use $\omega=2\pi f$.
5. Cancel the common u in the parentheses.
6. Convert from natural to common logarithms.
7. Coalesce 0.05 with the logarithmic term by use of conductor geometric mean radius $GMR=r/0.78$.
8. Transform to reactance: $X_{LL}=E_{LL}/I$.

Through these steps we obtain

$$X_{LL}=\frac{7.68f\times10^{-6}}{n}\times 0.46\log\frac{a_2a_4\ldots.a_n}{0.78ra_3a_5\ldots.a_{n-1}} \qquad (5)$$

This cumbersome form was found (as shown in the appendix) to reduce to

$$X_{LL}=\frac{3.53f\times10^{-6}}{n}\times \log\frac{4R}{0.78nr} \text{ ohms per 1,000 feet} \qquad (6)$$

This may be stated in terms of basic dimensions by substituting the value of R from equation 3. Then

$$X_{LL}=\frac{3.53f\cdot10^{-6}}{n}\log\frac{2(2r+t)}{0.78r} \qquad (7)$$

or, since

$$r=\sqrt{\frac{2CM/n}{2{,}000}} \qquad (8)$$

$$X_{LL}=\frac{3.53f\times10^{-6}}{n}\log 5.1\left(1+\frac{1{,}000t}{\sqrt{2CM/n}}\right)$$

ohms line-to-line per 1,000 feet (9)

where

f=frequency in cycles per second

t=thickness of insulation between ropes

n=the total number of ropes in the cable

CM=circular mils of conductor in each electric side of the cable; that is, half the total copper in the cable

Equation 9 shows that the reactance is nearly inversely proportional to the number of ropes per conductor, and only slightly dependent upon conductor dimensions.

This relationship is illustrated in Figure 4 at 10 kc for one thickness of separator, a range of common cable sizes, and two to six ropes per side.

Resistance

No calculations have been made on skin effect or proximity effect of the array of conductors under discussion here. It is evident that the breaking up of the conductor and the interaction between ropes will produce a better distribution of current than in ordinary conductors, and so decrease the resistance.

Tests have been made on 6-rope cables at high power levels with direct current and at 60, 890, and 9,600 cycles. The latter were rather crude and are included only because they furnish an "order of magnitude" value. More accurate tests are planned at various frequencies and with different cable sizes and numbers of ropes.

From available tests, the approximate ratio of a-c resistance to d-c values are shown in Figure 5.

Capacitance

The capacitance between the two electric sides of the cable depends upon whether it is dry or filled with water. An estimate is made most easily under the latter conditions, since the high dielectric constant of the water permits the first approximate assumption that the voltage stress is distributed uniformly across the rubber separator whose perimeter is p.

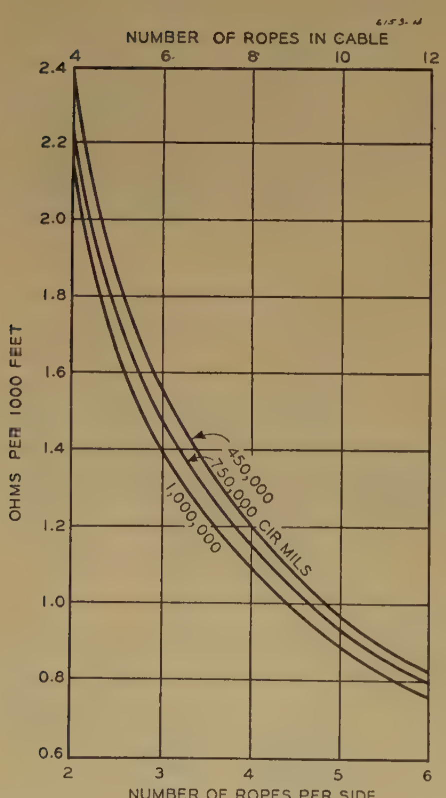

Figure 4. Inductive reactance at 10 kc versus number of ropes in three sizes of interleaved cable

This is the equivalent of a plate capacit[...] of width p and of length equal to the lengt[...] of the cable.

Since the perimeter of the separator [...] equal to that of insulators completely e[...] circling half the ropes, $p=2\pi nr/2=\pi n$[...] the capacitance, therefore, is

$C=0.224kp/t=0.224k\pi nr/t$ micromicrofarads per inch length (1[...]

where

t=thickness of the separator

k=specific inductive capacitance of t[...] separator

Taking $k=5$ for a rubber insulator, usi[...] a 1/16-inch thickness, changing to o[...] foot of cable length, and substituting t[...] value of r from equation 8, we get

$C=0.34\sqrt{n}\sqrt{2CM}$ micromicrofarads p[...] foot of cable length

Considering this as a lumped capacitanc[...]

$$X_c=\frac{1}{2\pi fc}=\frac{1}{2\pi f\times0.34\sqrt{n}\sqrt{2CM}}=\frac{0.47\times10^6}{f\sqrt{n}\sqrt{2CM}} \text{ megohm-feet of cable}$$

(1[...]

Figure 6 gives reactances calculate[...] from this formula.

An idea of charging current is obtaine[...] by taking 500 volts across a 10-foot lengt[...] of 6-rope 450,000-circular-mil cable [...] 10 kc:

$$I_c=\frac{500\times10}{20{,}000}=0.25 \text{ ampere}$$

Although much greater than the charg[...]ing current for ordinary cables, this is sti[...] negligible. Tests indicate that this curren[...] value is reduced by about ten in the dr[...] cable.

Leakage Losses

There are two causes of loss:

1. Current through the water when th[...] cable is water cooled.
2. Dielectric losses in the rubber insulatio[...]

In the standard cable, the current has [...] direct path through the water in th[...] "caps" of the two terminals. The d-c re[...]sistance of these paths in parallel is abou[...] 6,000 ohms in good city water. This re[...]sistance can be increased by special con[...]struction if required. In dry cable th[...] leakage resistance of a 6-foot long as[...]sembly is about 300,000 ohms.

The dielectric losses of insulation de[...]pend, of course, upon the material used[...] Neglecting losses in the water, a reason[...]ably high value for present inferior insula[...]

tion, is ten per cent of the charging current which has been derived previously:

$$\text{Dielectric loss} = \frac{0.10E^2}{X_c} \qquad (12)$$

For example, in the typical case of a 10-foot 6-rope 450,000-circular-mil cable, the dielectric loss at 500 volts, 10 kc will be

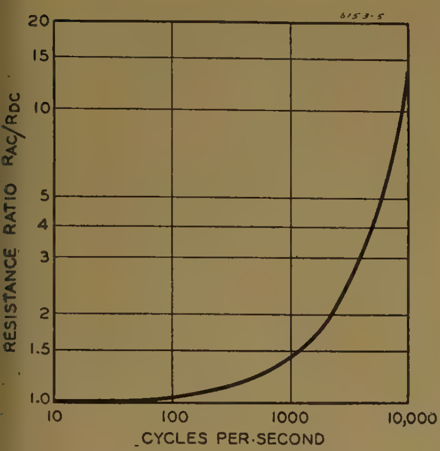

Figure 5. Approximate resistance ratio versus frequency for 6-rope interleaved cable

about 0.10 × 500 × 500/20,000 = 12 watts.

These losses are small enough so they can be neglected in most cases.

Conclusion

The electrical impedances of water cooled interleaved cables in the range of frequencies up to 10 kc are approximately

$$Z_{series} = \text{resistance ratio} \times \frac{23L}{CM} + j\frac{3.53f\times 10^{-9}L}{n} \times \log 5.1\left(1+\frac{1{,}000t}{\sqrt{2CM/n}}\right)\text{ohms} \qquad (13)$$

$$Z_{shunt} = 6{,}000 - j\frac{0.47\times 10^{12}}{fL\sqrt{n}\sqrt{2CM}}\text{ ohms} \qquad (14)$$

where

L = cable length in feet
CM = circular mils of conductor in each electrical side of cable
f = frequency in cycles per second
n = total number of ropes in the cable
t = thickness of separator in inches

As an example, when operating a 10-foot length of 6-rope 450,000-circular-mil cable at 10 kc,

$Z_{series} = 0.0064 + j0.0155$,
$Z_{shunt} = 6{,}000 - j2{,}000$

If carrying 2,000 amperes at 500 volts, voltage drop = $12.8 + j\,31 = 33.6/\underline{168°}$, and leakage current = $0.1 + j\,0.2$

In this frequency range, leakage is unimportant, and estimates may be made on the basis of equation 13. This is borne out by installations at 900 cycles and at 10,000 cycles where production economy demanded a flexible conductor.

Any modification from the commercial 6-rope cables, that are considered for the purpose of decreasing the voltage drop, must be viewed in the light of decreased flexibility because of larger diameter, and increased cost as a result of more difficult and nonstandard manufacture.

Appendix

Another method of deriving equation 6 is through geometric means, that is,

$$L_t = 2\,ln\frac{G_{to}}{G_{tt}} \qquad (15)$$

where

L_t = inductance associated with the inner ropes
G_{to} = geometric mean distance between all the inner ropes and all the outer ones
G_{tt} = geometric mean distance of the inner ropes with respect to themselves

To state the G's in terms of cable dimensions the simplest approach, suggested by H. B. Dwight, is to apply Guye's theorem which states that for m points or circles spaced equally around a circular line of radius R, the geometric mean distance of one from the others is

$$g = Rm^{\frac{1}{m-1}} \qquad (16)$$

(In this development, g is used for the distance from one rope to others, G denoting the distance between a group of conductors and others.)

First consider the inner ropes, letting m equal the number of inner ropes. Then the geometric mean distance from one inner rope to the other inner ropes is

$$g_{to} = Rm^{\frac{1}{m-1}} \qquad (17)$$

To find the geometric mean distance of the complete array with respect to itself, we also refer to the well-known relation that for a circle

$$g_{ts} = r\epsilon^{-1/4} = 0.78r = r' \qquad (18)$$

Then the geometric mean distance of the inner set of ropes is obtained from

$$G_{tt}^{\,m} = g_{ts}g_{tt}^{\,m-1} \qquad (19)$$

$$= r'\left(Rm^{\frac{1}{m-1}}\right)^{m-1}$$

$$= mr'R^{m-1} \qquad (20)$$

Having found the "self" geometric mean distance G_{tt}, we proceed to the "mutual" geometric mean distance G_{to} between inner and outer wires. This is done indirectly (since Guye's theorem gives a direct solution for arrays of equidistant points like . . *iii*. or . . *ioio* . ., but not for . . *ooioo* . . .) by finding $g_{t(t+o)}$, the geometric mean distance from i to all other ropes, and then combining that with g_{tt} to obtain G_{to} from equation 22.

That is, using equation 16 with $2m$ = total number of ropes,

$$g_{t(t+o)} = R(2m)^{\frac{1}{2m-1}} \qquad (21)$$

Now because of the symmetry of the array, $g_{t(t+o)}$ also could be obtained by the process

$$g_{t(t+o)} = \sqrt[2m-1]{g_{tt}^{\,m-1}G_{to}^{\,m}} \qquad (22)$$

which gives

$$G_{to}^{\,m} = g_{t(t+o)}^{\,2m-1}/g_{tt}^{\,m-1} \qquad (23)$$

where G_{to} is the geometric mean distance of all the inner ropes with respect to all the outer ones. Substituting equations 17 and 21 in equation 23,

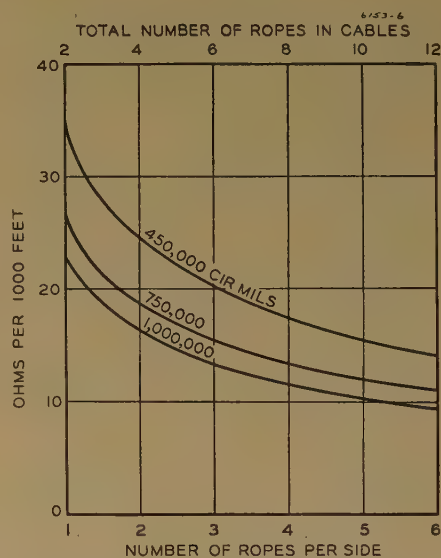

Figure 6. Calculated capacitive reactance at 10 kc versus number of ropes in three sizes of interleaved cable

$$G_{to}^{\,m} = R(2m)^{\frac{1}{2m-1}}/Rm^{\frac{1}{m-1}}$$

$$= 2mR^{2m-1}/mR^{m-1}$$

$$= 2R^m \qquad (24)$$

Combining equations 24 and 20,

$$\left(\frac{G_{to}}{G_{tt}}\right)^m = \frac{2R^m}{mr'R^{m-1}}$$

$$= \frac{2R}{mr'} \qquad (25)$$

$$m\cdot ln\frac{G_{to}}{G_{tt}} = ln\frac{2R}{mr'} \qquad (26)$$

Since $m = n/2$,

$$ln\frac{G_{to}}{G_{tt}} = \frac{2}{n}ln\frac{4R}{mr'} = \frac{2}{n}ln\frac{4R}{0.78mr} \qquad (27)$$

Influence of Magnetic Materials on the Welding Characteristics of Resistance Welding Machines

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MOST MEN associated with resistance welding machines are conscious of the fact that the introduction of magnetic material into the throat of such equipment causes a reduction of the welding current. The degree of reduction and simple mathematical tools for estimating it are contained in this paper.

Steel sheets and products fabricated from them constitute the majority of welding applications. It was of interest to determine the effect on the welding current by changing

1. The thickness of the sheets.
2. The length inserted into the welder throat.
3. The position in the welder throat.
4. The width of the sheets.
5. The actual current used for welding.

All of these factors could react to reduce the welding current, some more severely than the others.

The change in welding current was evaluated by measuring the reactive and resistive voltage components caused by the steel sheets in the welder throat.

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These voltage components in combination with the short-circuit impedance constants of the welding machine provide the necessary information.

Consequently, with a knowledge of these impedance constants and the nomogram indicating the effect of steel, it is a brief routine operation to calculate the change in welding current caused by the insertion of magnetic material into the secondary welder loop.

The apparent loss of current arises from the more intense magnetic field or increase in flux linkages and from the core loss resulting as heat generated in the steel sheets, a natural consequence of being subjected to an alternating magnetic field. Some effort was directed toward determining the magnitude and distribution of these flux linkages so that a better basic understanding of the influence of magnetic materials could be realized.

The magnitude of the magnetic field produced in the welder throat will depend primarily on the welding current. It is realized that a wide range of welding current can be used to make spot welds in mild or low carbon steel. The desirable maximum and minimum points of this range, together with other associated welding variables, are given in Table I. Figure 1 summarizes the welding variables for seam welding of low carbon steel. Figure 2 lists the welding variables fo spot welding of austenitic stainless steels The data shown in both Figures 1 and are the tentative recommended prac tices of the American Welding Society.

These combinations of welding vari ables produce satisfactory spot welds an seam welds, and justify their use for fu ther work of the influence of magneti materials.

Apparatus

The welding was performed on a Re sistance-Welder-Manufacturers-Associa tion size-2 press welder manufacture by the Taylor-Winfield Corporation an capable of being used as a spot and pro jection welder. A cross sectional vie along the length of the welder throat shown in Figure 3. Area and positio of the current carrying members var at different lengths in the welder throa

Procedure

It was desirable to make the necessar electrical measurements while makin actual welds with the steel sheets in serted in the welder throat. Howeve the slight distortion of the current wav form caused by the volume of steel adde to the circuit and the attempt to accu rately read the small changes in power fac tor from a watt-galvanometer oscillo graphic trace made it imperative to sepa rate the resistance caused by making weld from the resistive component cause by the presence of steel sheets. It wa evident that the welding resistance R was much larger than the resistive com ponent R_{rs}, especially in thin sheets.

MEASUREMENT OF WELDING RESISTANC R_w

The welding resistance R_w is the resist ance from electrode to electrode acros

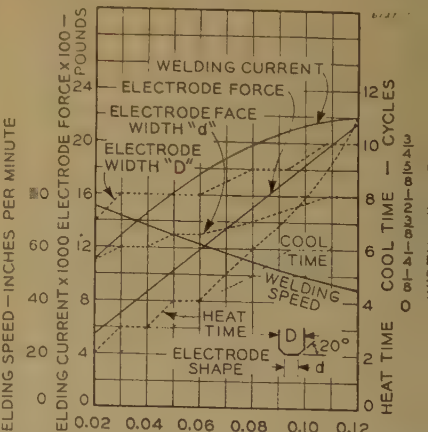

Figure 1. Low carbon steel seam weldin data

Using equation 26 in equation 15, we find that

$$L_i = \frac{4}{n} ln \frac{4R}{0.78nr} \text{ abhenries per centimeter} \quad (28)$$

Appropriate changes to practical units and conversion to reactance produce equation 6.

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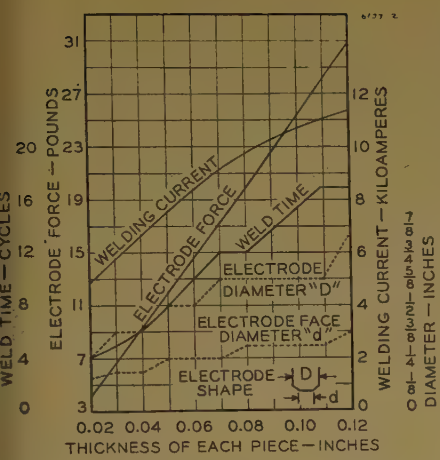


Figure 2. Austenitic stainless steel spot welding data

the work pieces when welding small coupons. It is the resistance added to the electrical circuit by welding operation itself and is independent of steel that extends into the welder secondary loop. R_w differs from the resistance R used in making the weld by that energy or heat that is dissipated into the work pieces and at the electrode faces.

Expressed mathematically

$$\text{Energy for welding} = I^2Rt + \text{losses}$$
$$= I^2R_wt$$

where

I = welding current
t = time of current flow
R = resistance used for welding
R_w = effective resistance present to the flow of welding current caused by welding

The value of R_w is the important quantity in welder operational characteristics, and it is fortunate that it can be measured easily while R is impossible to determine accurately.

The resistance R_w was measured when making a weld in small coupons by taking the voltage drop across the two electrodes at two points as close to the weld as possible and dividing this voltage by the secondary current. When welding stainless steel, the drop from the top of one sheet to the bottom of the other one (sheet to sheet resistance, V_2) also was measured. A schematic diagram of the setup is shown in Figure 4. The inductive pickup was minimized by carefully twisting the voltage leads together and holding them perpendicular to the electrodes.

Simultaneous oscillographic records were taken of the voltage drops, primary current, and watts. The primary current was converted into the secondary or welding current by multiplying by the turns ratio of the welding transformer. The magnetizing current was neglected being less than two per cent in the most exaggerated case.

The variations of R_w with time during spot welding for several thicknesses of low carbon steel with different combinations of welding variables is shown in Figures 5 and 6. A less comprehensive picture of the variation of R_w with time for austenitic stainless steel is shown in Figure 7. A similar set of data for seam welding low carbon steel is presented in Figure 8. By selecting average values of R_w for spot welding low carbon steel and stainless steel under the conditions specified by Table I and Figure 2, the variation of R_w with sheet thickness results as shown in Figure 9.

The mild steel used in this investigation was covered with a slight film of oil to prevent rusting, being previously hot rolled and pickled. Before welding the steel was wiped with a clean rag to remove any dirt that accumulated during storage. The surface was still perceptibly oily after wiping. This type of surface condition, most often used in production applications, increases R_w beyond that expected from degreased sheets.

Measurement of Resistive Voltage Component, V_{rs} and Reactive Voltage Component, V_{xs} Caused by the Insertion of Magnetic Material or Steel into the Welder Throat

These voltage increments were measured without involving the welding resistance R_w. A hole two inches in diameter with its center two inches from the front edge and midway across the width was drilled in two equal thicknesses of steel sheet (Figure 10). The sheets were insulated from each other by a sheet of

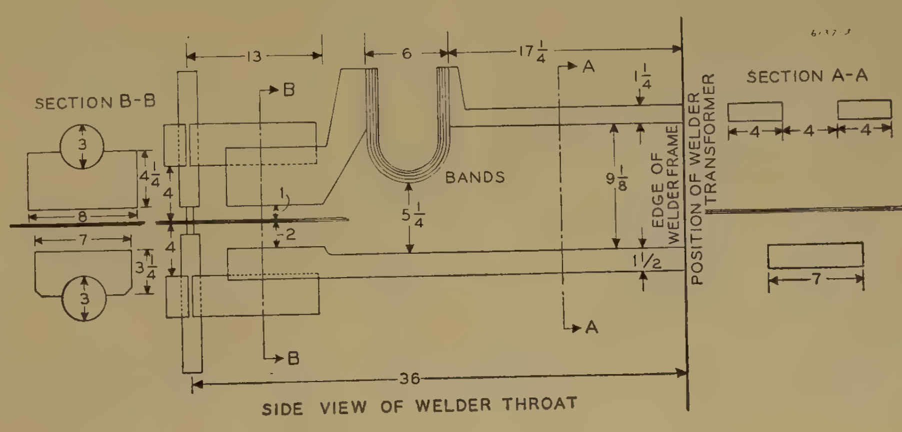


Figure 3. Side view of welder throat construction showing copper and copper-alloy current conducting members

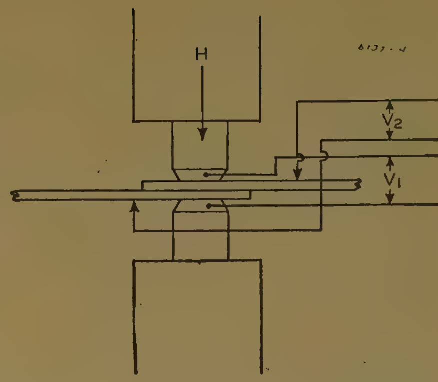


Figure 4. Electrode setup for measurement of welding resistance and sheet-to-sheet resistance

$$\text{Welding resistance } R_w = \frac{V_1}{I}$$

$$\text{Sheet resistance } R_{ss} = \frac{V_2}{I}$$

paper, inserted in the welder throat midway between the welder horns, and locked in position with wooden blocks in order to counteract the magnetic force when the welder current was applied. A water-cooled copper tube was inserted through the hole in the sheets and connected to the welder horns. With the steel in place, the current necessary to make a weld and form an equivalent magnetic field was passed through the secondary circuit of the welder. A magnetic contactor, manually controlled, allowed interruption of the current whenever desired. The frequency of the applied voltage was 60 cycles per second.

Electrical values were measured using instruments equipped with pointer stops and capable of reading the rms values of distorted wave forms. Readings were taken of primary voltage, primary current, primary power, welder transformer turns ratio, and depth of steel in welder throat.

The steel sheets were prepared initially to extend the maximum depth into the welder throat. After each set of measurements six inches would be sheared off the

Table I. Low Carbon Steel Spot Welding Data

Data Common to All Classes of Spot Welds						Welding Variables for Best Quality Class-A Weld					Welding Variables for Better Quality Class-B Weld					Welding Variables for Good Quality Class-C Weld				
Thickness of Thinner Outside Piece		Electrode Diam and Shape (See Figure 1)					Min Weld Strength and Electrode Force					Min Weld Strength and Electrode Force					Min Weld Strength and Electrode Force			
T, In.	United States Standard Gauge	Max ρ, In.	Min D, In.	Min Weld Spacing, Center Line to Center Line, In.	Min Contacting Overlap (Note 8), In.	Weld Time Single Impulse, Cycles	Force, Lbs	Load, Lbs	Approx. Current, Amps	Approx. Diam of Fused Zone, In.	Weld Time, (Note 8), Cycles	Force, Lbs	Load, Lbs	Approx. Current, Amps	Approx. Diam of Fused Zone, In.	Weld Time (Note 7), Cycles	Force, Lbs	Load, Lbs	Approx. Current, Amps	Approx. Diam of Fused Zone, In.
0.010	32	1/8	3/8	1/4	3/8	4	200	200	4,000	0.13	5	130	170	3,700	0.12	15	65	130	3,000	0.11
0.021	25	3/16	3/8	3/8	7/16	6	300	450	6,100	0.17	10	200	390	5,100	0.16	22	100	320	3,800	0.14
0.031	22	3/16	3/8	1/2	7/16	8	400	840	8,000	0.21	15	275	770	6,300	0.20	29	135	680	4,700	0.18
0.043	19	1/4	1/2	3/4	1/2	10	500	1,300	9,500	0.23	21	360	1,160	7,500	0.22	38	180	1,050	5,600	0.21
0.049	18	1/4	1/2	7/8	9/16	12	650	1,550	10,300	0.25	24	410	1,420	8,000	0.23	42	205	1,320	6,100	0.22
0.061	16	1/4	1/2	1 1/16	5/8	14	800	2,050	11,600	0.27	29	500	1,900	9,000	0.26	48	250	1,770	6,800	0.25
0.077	14	5/16	5/8	1 3/8	11/16	21	1,100	2,800	13,300	0.31	36	650	2,620	10,400	0.30	58	325	2,420	7,900	0.28
0.092	13	5/16	5/8	1 5/8	3/4	25	1,300	3,550	14,700	0.34	44	790	3,450	11,400	0.33	66	390	3,160	8,800	0.31
0.107	12	3/8	5/8	1 13/16	13/16	29	1,600	4,600	16,100	0.37	50	960	4,350	12,200	0.36	72	480	4,050	9,500	0.35
0.123	11	3/8	5/8	2	7/8	30	1,800	6,000	17,500	0.40	60	1,140	5,550	12,900	0.39	78	570	5,200	10,000	0.37

Notes:

1. Type of steel SAE 1010.
2. Material free from grease, scale, and dirt.
3. Welding conditions determined by thickness of thinnest outside piece *T*.
4. Data for total thickness of pileup not exceeding 4*T*.
5. Minimum spacing is that spacing for two pieces for which no special precautions are necessary to compensate for shunting of adjacent welds. For 3 pieces increase spacing 30 per cent.
6. Electrode material, class 2, minimum conductivity, 75 per cent copper; minimum hardness, 75 per cent rockwell *B*.
7. Weld times from 1 to 30 cycles consisted of a single impulse, from 32 to 60 cycles consisted of two equal impulses, and from 63 to 90 cycles consisted of three equal impulses.
8.

previous length until five inches extended into the welder throat.

The short-circuit constants or welder electrical constants without any steel in the welder throat but with the proper horn spacing were checked after each set of measurements. The resulting values of impedance, reactance, and resistance referred to the secondary side are designated as Z_0, X_0, and R_0, respectively.

The values of the voltage components caused by the addition of steel in the welder throat were determined in the conventional manner (Appendix I). The voltages V_{rs} and V_{xs} are the resistive and reactive components referred to secondary circuit.

These components are plotted for several thicknesses of sheet and different horn spacings in Figure 11. The components were the same for 8- and 16-inch horn spacing within the accuracy of measurement. The effect of sheet width on the voltage components is shown in Figure 12. The change of resistive and reactive voltage components with welding current for one gauge of sheet is illustrated in Figure 13.

In order to establish the extent to which cold-working the austenitic stainless steel made it magnetic because of the formation of small areas of the metallurgical constituent ferrite, two sheets of the largest thickness available were checked at various dimensions in the welder throat. Values of V_{rs} and V_{xs} are shown in Figure 14.

Measurement of Flux Patterns

While the absolute values of resistive and reactive voltage components caused by steel in the welder throat are the working tools in electrical characteristics calculations, a study of the changes in flux and flux pattern in the welder throat circuit is most beneficial to understanding the phenomenon and applying an engineering analysis to the problem.

The flux was measured by evaluating it from the voltage induced in a search coil. The method used for calculating flux from the induced voltage is described in Appendix III. In all tests two equal thicknesses of steel sheet 12 inches long were placed in the welder throat so that the front edge was two inches from the edge of the shorting bar, and the width was distributed equally on each side of the horn center line. The influence of the width of the sheet on the variation of maximum flux density along width of the sheet is shown in Figure 15.

As saturation of the low carbon steel sheets would occur at flux densities of approximately 125,000 lines per square inch, it is obvious that the normal welding current saturated the sheets at the center, or midway between the current carrying conductors. In order to determine how much the welding current could be decreased before the steel directly between the conductors would not saturate, two of the thickest sheets used in past work were subjected to different currents and the maximum flux density measured

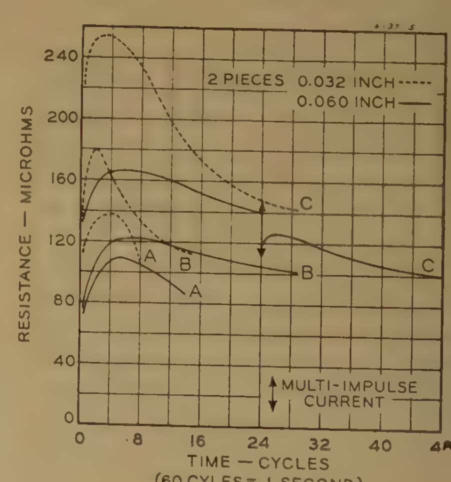


Figure 5. Variation of welding resistance with time in spot welding low carbon steel sheet

A, B, and *C* refer to classes of welds

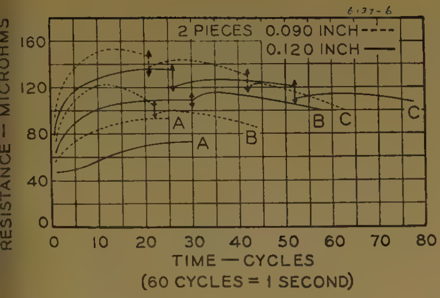


Figure 6. Variation of welding resistance with time in spot welding low carbon steel sheet

A, B, and *C* refer to classes of welds

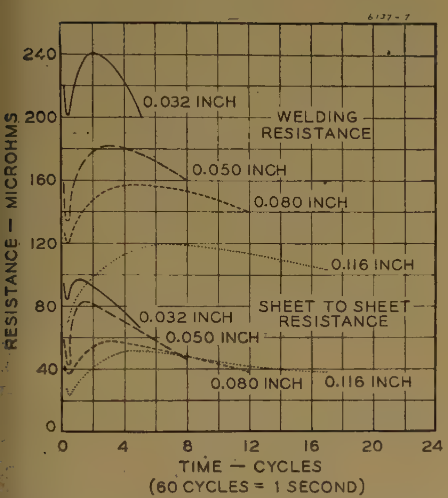


Figure 7. Variation of welding resistance and sheet-to-sheet resistance with time in spot welding austenitic stainless steel sheet of half-hard temper

(Figure 16). The thickest sheets were selected for this test as the ratio of welding current to sheet thickness decreases with increasing sheet thickness. Saturation still exists at the lowest value of current selected

Results

Welding Resistance R_w

The welding resistance R_w consists of two main components

1. The resistance from electrode to sheet.
2. The resistance from sheet to sheet.

The first component consists primarily of contact resistances. The second component consists of a contact resistance between the sheets and the resistance of the sheets themselves. From the test results on spot welding stainless steel in which these components were measured individually, it is notable that the resistance from sheet to sheet is approximately only 40 per cent of the total. It is expected that a similar relationship also would prevail for mild steel. Some past investigations[2-4] have indicated this approximate relationship. It is evident that the contact resistance between electrode and work piece is an appreciable part of the welding resistance and causes considerable heat generation at the electrode faces. Such being the case, it is very essential to cool the electrodes adequately in spot welding to prevent excessive deformation and to have the data apply.

While it is instructive to determine the components of the welding resistance R_w individually, the basic effect on the electric circuit is summarized in the value of R_w. From past experience it would be anticipated that R_w would be a function of the current density at the electrode face and the force density or pressure at the electrode face. As the variation of R_w with time could not be related satisfactorily to these two factors, average values of R_w were selected from Figures 5 and 6. Using these values and the associated data contained in Table IV, the following empirical relationship was determined for spot welding low carbon steel:

$$R_w = 333 \frac{(D^{1.5}+1.7)}{P+2}$$

where

R_w = welding resistance in microhms
D = current density in (amperes per square inch) $\times 10^{-5}$
P = pressure in (pounds per square inch) $\times 10^{-3}$

The agreement between the calculated and measured values of R_w also is shown in Table IV. The equation gives passable results and point out that the heat generated in welding increases more rapidly than the square of the current, and reduction in electrode force causes appreciable increase in welding resistance.

In the empirical equation for R_w, the resistivity of the steel is not shown explicitly. However, it is erroneous to assume that the resistivity has no effect on welding resistance. In spot welding stainless steel, the values of R_w are considerably larger than those for mild steel of like gauges. The difference in resistivity,

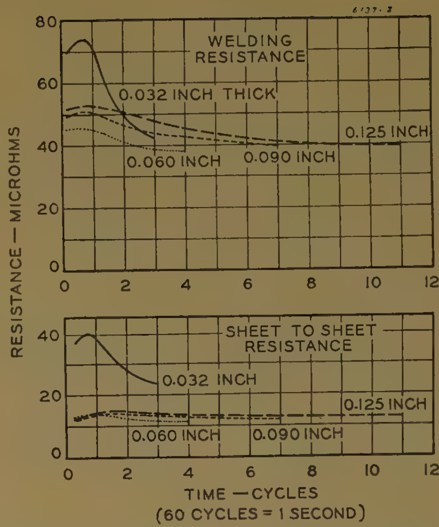


Figure 8. Variation of welding resistance with time in seam-welding low-carbon steel sheet

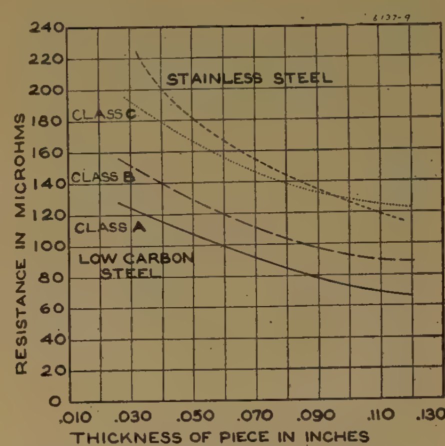


Figure 9. Variation of average welding resistance with thickness of sheet in spot welding

12.3 and 73 microhms per centimeter cube, and surface condition between commercial hot rolled pickled and oiled mild steel and bright finished stainless steel, account for the marked differences in welding resistance between these materials.

The values of welding resistance in seam welding should depend on the same factors as in spot welding but the different nature of the electrode contact area while welding and the shunting effect of

Table II. Materials Used in Welding Investigation

Type of Steel	Approximate Chemical Composition C	Si	Mn	Cr	Ni	P	S	Ultimate Tensile Strength, Lbs Per Sq In.
Commercial-quality cold-rolled steel sheet	0.08	0.34	0.35			0.01	0.01	45,000
Austenitic stainless steel, half-hard temper	0.10	0.50	1.1	17.3	7.3	0.015	0.006	160,000

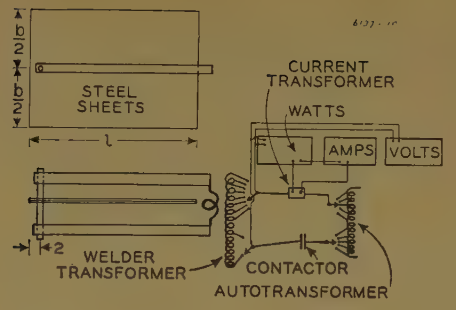


Figure 10. Circuit for measuring impedance of steel work pieces inserted in welder throat

the previous welds in a pressure tight seam cause deviations from the values of R_w measured in spot welding. It was rather surprising to discover that both the welding resistance R_w and the sheet to sheet resistance R_{ss} were almost equal in the range of thicknesses from 0.060 to 0.120 inch.

The emphasis on the determination of welding resistance was justified entirely as the resistive voltage component caused by it is of much greater magnitude than the resistive voltage component V_{rs} caused by the eddy current loss in the steel sheets inserted into the welder throat. If this factor were assumed carelessly, the final application of the data of V_{rs} would have given avoidable erroneous results. The values of R_w are so high in some cases that many resistance welders in operation today probably are functioning at higher power factors than normally assumed.

The Reactive Voltage Component V_{xs} and the Resistive Voltage Component V_{rs} Caused by the Presence of Steel in the Welder Throat

Engineers making electric circuit calculations are familiar with the procedure of considering a resistance welder secondary loop in terms of reactive and resistive impedance constants. These are combined with impedance values evaluating the welding effect and the totals placed as the loaded secondary circuit of the welder transformer. This procedure and much factual data on welder secondary circuits are contained in an article by J. H. Cooper.[5] This practice must be altered when evaluating the presence of steel in the welder throat as ohmic values of reactance and resistance. These impedances change with the welding current. The effect of steel in the throat is to produce reactive and resistive voltage components opposing the applied secondary voltage of the transformer. These voltage components are almost independent of current over the normal welding range (Figure 14). Thus

$$V_{xs}=IX_s=C_1 \quad V_{rs}=IR_s=C_2$$

Consequently, if ohmic values of reactance and resistance are desired to express the presence of steel in the welder throat, the values must be selected in combination with the welding current to give the correct voltage drop.

Saturation of the steel sheets by the welding current is the prime reason why V_{xs} and V_{rs} remain almost independent of current. An elementary study of the flux changes in the steel sheets is instructive in pointing out the limitations of the data.

The magnetic field in air between two parallel conductors is shown in Figure 17. This field will be distorted when steel sheets are added, but the flux still will concentrate at the section of steel directly between the conductors. It must be realized that the effect of the steel sheets in distorting the air flux field as well as the effect of causing more flux linkages have been combined to evaluate V_{xs} and V_{rs}. In other words, the values of welder shortcircuit impedance constants X_o and R_o have been assumed constant in spite of the air field distortion and the higher harmonics in the welding current caused by the presence of steel sheets in the welder throat. This assumption is valid providing V_{xs} and V_{rs} do not become too large a part of the applied secondary voltage as will be the case in most welder secondary circuit construction.

The steel sheets are saturated by the magnetizing force caused by the welding current. The relationship of welding current, flux, induced voltage, and the flux-current curve for a specific thickness and length of steel are shown diagrammatically in Figure 18. As the welding current increases, the flux hardly increases as the intrinsic flux density B_i-$(B-\mu_oH=B_i)$ has reached its limiting value.* The area of the induced voltage curve is a measure of the maximum flux, and as the maximum flux under saturated conditions will be almost constant the average value of induced voltage will be al-

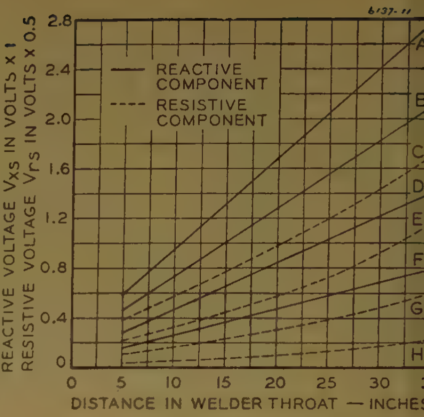


Figure 11. Variation of reactive and resistiv voltage components caused by the presence steel in the welder throat with distance throat

A and *C*. Two pieces 0.119 inch thic 18,000 amperes

B and *E*. Two pieces 0.090 inch thic 15,000 amperes

D and *G*. Two pieces 0.060 inch thic 12,000 amperes

F and *H*. Two pieces 0.032 inch thic 8,300 amperes

Duplicate results for 8-and 16-inch horn spac ings

most constant. The effect of change c welding current should be to change th wave form of flux and the wave form c the induced voltage. In circuit calcula tions it is the effective value and not th average value of induced voltage tha must be investigated. The measurement indicated that both the effective and av erage values of induced voltage staye almost constant, in spite of change o welding current. In order to have thi condition prevail with such wave form of induced voltage as recorded, it is nec essary to have the flux current relation ship change with different magnetizin forces.

The value of V_{xs} is proportional to th increase in length of steel in the welde throat. Under saturated conditions th maximum flux density is constant, an the increase in induced voltage is the re sult of the increase in flux caused by th increase in cross-sectional area provide by the additional length for a specifi thickness.

Horn spacing or the distance betwee the throat bars does not influence V_x appreciably. The steel sheets becom saturated at any specific position in almos the same length of time from the zer point of the current wave. There is n

Table III. Impedance of Welding Machine and Welding Transformer at a Frequency of 60 Cycles Per Second

Horn Spacing, Inches	Throat Depth, Inches	Impedance Z_o, Microhms	Reactance X_o, Microhms	Resistance R_o, Microhms
8	36	273	268	54
11	36	344	340	56

Note: There is no measurable change in impedance constants at secondary currents of 10,000 and 20,000 amperes.

* The intrinsic flux density B_i is a measure of tha part of the flux density attributable to the ferro magnetic characteristic of the material. When th material becomes saturated, it can make no furthe contribution to the flux density. As the mag netizing force H is increased beyond the valu which saturates the material, the slope of th $B-H$ curve becomes μ_o, the permeability of fre space, and B_i becomes constant.

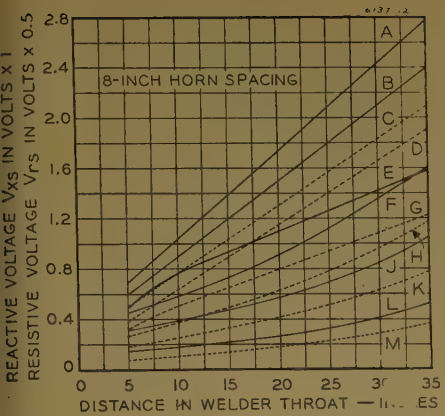


Figure 12. Variation of reactive and resistive voltage components with distance in throat and width of sheet

A, B, and *E.* Reactive voltages for two pieces 0.118 inch thick and widths of 36, 18, and 9 inches, respectively

F, J, and *L.* Resistive voltages under the above conditions

C, D, and *G.* Reactive voltages for two pieces 0.090 inch thick and widths of 36, 18, and 9 inches, respectively

H, K, and *M.* Resistive voltages under the same conditions

need to have the sheets equally spaced between welder horns to have the data apply.

It might be anticipated that V_{xs} would be independent of the width of the steel sheet if the steel were saturated at the section under the center line of the welder horns and thus limited the flux. By considering the air flux pattern, it is clear that narrow sheets will coerce less flux lines to enter the sheet than wide ones. As the current and magnetizing force change with time in the normal periodic manner, the flux will follow a similar pattern. As the saturation density is reached the preference for flux to pass into the sheet is eliminated. However, with narrow sheets it takes a longer period of time of a half-cycle of current flow to reach the saturation density. The resulting induced voltage is smaller as the rate of change of flux with time is smaller. It would be anticipated that after a certain width of sheet is reached, greater widths would cause negligible increase. This factor is associated with the distance between throat conductors or horn spacing and shape of conductors. The correction factors in Figure 20 are based on 8-inch horn spacing.

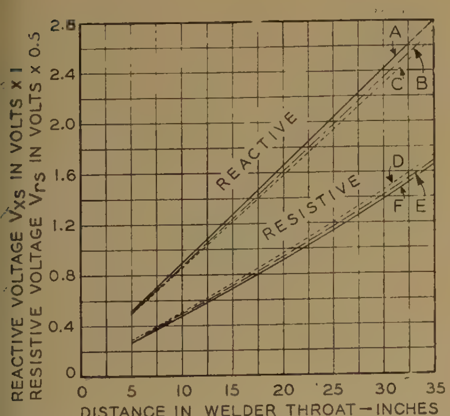


Figure 13. Variation of reactive and resistive voltage components with distance in throat and welding current. Two pieces of low carbon steel 0.118 inch thick and 36 inches wide

A and *F.* 19,000 amperes
B and E. 15,000 amperes
C and *D.* 10,000 amperes

Austenitic stainless steel is considered nonmagnetic, but cold working the steel causes the formation of small areas of iron or ferrite. These areas of ferrite would certainly reach saturation under the magnetizing force caused by the welding current. However, for the half-hard condition, the resulting values of V_{xs} and V_{rs} are small. The full-hard condition would have raised V_{xs} and V_{rs}. A word of caution is advisable in considering the magnetic effect in stainless steel. Those which are austenitic can be considered nonmagnetic for all practical purposes. However, those which are ferritic will be magnetic and should be considered having effects similar to mild steel.

The resistive voltage component multiplied by the secondary current represents the real power or core loss in the steel sheets. The great thickness of the sheet implies that the principle loss will be in the form of eddy currents. In the measurements, one steel sheet was insulated from the other, which means that the eddy currents would reach their maximum value near the outer surfaces of the sheets. If the sheets were in such intimate contact that the contact resistance between them could be considered zero, the resistive voltage component would increase theoretically to four times its former value. In most welding conditions the first case is more likely to exist than the second one, although some discretion should be used in applying the data.

The resistive voltage component caused by the presence of steel in the welder throat is normally a small part of the total resistive component composed also of the welder component I_tR_o and the component finished by the welding resistance I_tR_w.

Application of the Data

After making an analysis of the data and the limitations imposed upon them, it was desirable to compile these data into usable form. The values of V_{xs} and V_{rs} in Figure 11 were examined. By drawing straight lines for the V_{rs} variation with length, the empirical formulas and the resulting nomogram shown in Figure 20 were derived. The value of V_{rs} does not include the effect of welding resistance. The correction for width was obtained from Figure 12. The correction for difference in welding current from the base current was obtained from Figure 13.

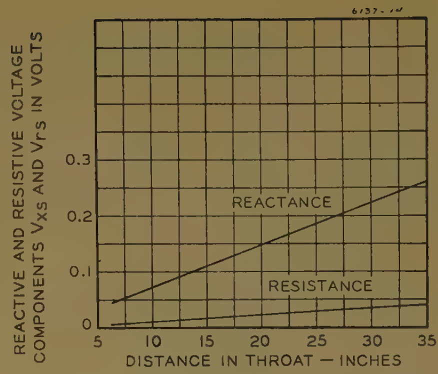


Figure 14. Variation of reactive and resistive voltage components with distance in throat for two pieces of austenitic stainless steel 0.109 inch thick, 24 inches wide and half-hard temper

The nomogram can be used for either spot welding or seam welding. For this reason, the effect of welding resistance is not included. The value of resistive voltage caused by the welding resistance can be selected from Figure 19. It is advisable to compare the desired welding technique with the technique outlined in Table I and Figures 1 and 2 so that a satisfactory choice of welding resistance voltage can be made.

Two examples of application of the data are shown in Appendix II.

Conclusions

1. The steel plates inserted in a welder throat will be saturated by the magnetic field produced by the welding current under normal conditions of operation.

2. The effect of the presence of steel plates (magnetic material) in the welder throat is measured best by expressing it as a reactive and a resistive voltage component. These voltage components are almost independent of welding current over the normal spot welding range.

3. The normal procedure of expressing the electric circuit of a welder in terms of resistance and reactance should be altered when the effect of steel plates is included. The circuit should be expressed in terms of reactive and resistive voltage drops, only some of which will be proportional to the current.

4. The effect of welding resistance and its

Table IV. Comparison of Calculated and Measured Values of Welding Resistance and Factors Affecting Welding Resistance

Thickness of Each Piece, Inches	Welding Current, Amperes	Electrode Force, Pounds	Force Density, Pounds Per Square Inch	Current Density, Kiloamperes Per Square Inch	Welding Resistance Rw, Microhms	
					Measured	Calculated
0.032	8,000	425	15,400	291	125	128
0.032	6,400	280	10,300	233	150	146
0.032	4,730	135	4,900	172	190	189
0.060	11,500	800	16,300	234	102	97
0.060	9,000	490	10,000	183	115	116
0.060	6,850	245	5,000	139	140	158
0.090	14,300	1,250	16,300	187	78	78
0.090	11,400	770	10,000	149	102	98
0.090	8,520	380	5,000	112	132	137
0.120	17,120	1,800	16,300	155	69	66
0.120	12,800	1,110	10,000	116	90	82
0.120	10,000	550	5,000	91	126	122

resulting voltage component is normally larger than the resistive voltage component caused by the presence of steel sheets for gauges less than 0.125 inch. As the thickness increases the effect of welding resistance voltage drop decreases and the resistive effect of the sheets becomes the predominating factor.

5. The effect of magnetic material in the welder throat is a function of the thickness and width of the sheets and of the secondary circuit constants of the welder. Resistance welders with large horn spacing are less influenced than welders with small horn spacing for the same throat depth. In other words, the larger the secondary voltage to supply a necessary welding current, the less the influence of the constant bucking voltage caused by the steel sheets.

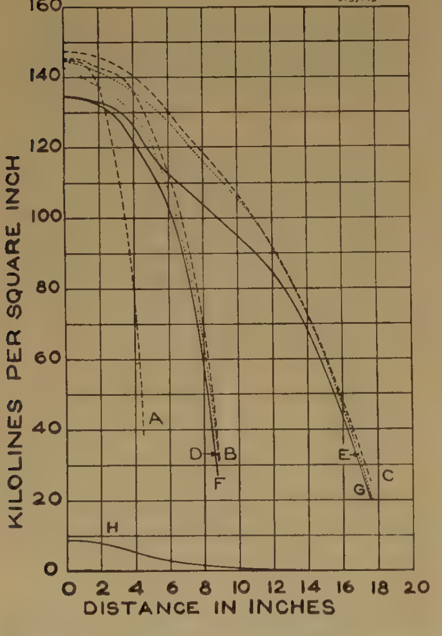


Figure 15. Variation of maximum flux density with distance from center line of electrodes at right angles to the welder throat

A, B, and *C.* Sheets 0.120 inch thick and 9, 18, and 36 inches wide, respectively, at 18,000 amperes

D and *E.* Sheets 0.090 inch thick and 18 and 36 inches wide, respectively, at 15,000 amperes

F and *G.* Sheets 0.032 inch thick and 18 and 36 inches wide, respectively, at 8,300 amperes

H. For magnetic field in air at 17,500 amperes

Appendix I. Determination of V_{rs} and V_{xs} from Instrument Readings

V_p = primary voltage, volts
I_p = primary current, amperes
W_p = primary power, watts
l = depth of steel in throat, inches
a = transformer turns ratio
V_t = secondary voltage, volts
I_t = secondary current, amperes
W_t = secondary power, watts
Z_p = impedance referred to primary winding, ohms
R_p = effective resistance referred to primary winding, ohms
X_p = reactance referred to primary winding, ohms
R_t = impedance referred to secondary winding, ohms
R_t = effective resistance referred to secondary winding, ohms
X_t = reactance referred to secondary winding, ohms
Z_s = impedance (for a specific current) of steel sheets referred to secondary winding, ohms
X_s = reactance (for a specific current) of steel sheets referred to secondary winding, ohms
R_s = resistance (for a specific current) of steel sheets referred to secondary winding, ohms
V_{rs} = resistive voltage component caused by steel sheets, volts
V_{xs} = reactive voltage component caused by steel sheets, volts
Z_0 = impedance of welder secondary circuit under short-circuit conditions, ohms
X_0 = reactance of welder secondary circuit under short-circuit conditions, ohms
R_0 = resistance of welder secondary circuit under short-circuit conditions, ohms

$$Z_p = V_p/I_p = a^2(Z_0+Z_s) \qquad (1)$$
$$Z_s = (Z_p/a^2) - Z_0 \qquad (2)$$

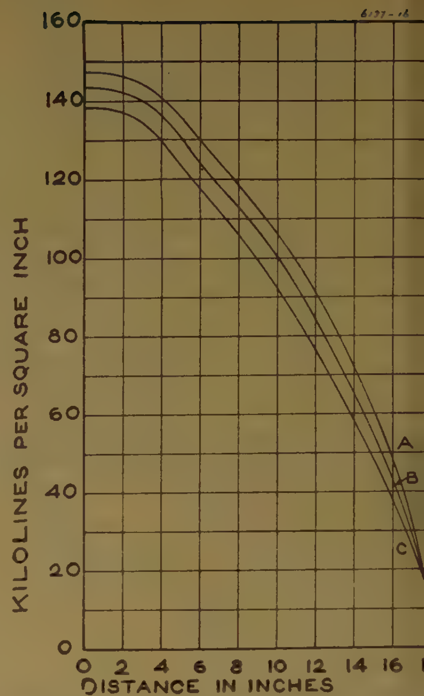


Figure 16. Variation of maximum flux densit with distance from center line of electrodes right angles to the welder throat and wi welding current for two pieces 0.120 in thick, 36 inches wide

A. 18,500 secondary amperes
B. 14,400 secondary amperes
C. 10,000 secondary amperes

$$R_p = W_p/I_p^2 = a^2(R_0+R_s) \qquad (3$$
$$R_s = (R_p/a^2) - R_0 \qquad (4$$
$$X_p = \sqrt{Z_p^2 - R_p^2} = a^2(X_0+X_s) \qquad (5$$
$$X_s = X_p/a^2 - X_0 \qquad (6$$
$$I_t = I_p(a) \qquad (7$$
$$V_{rs} = I_t R_s \qquad (8$$
$$X_{xs} = I_t X_s \qquad (9$$

The foregoing equations are used for th calculations of two sheets 0.118 inch thicl 18 inches wide, and inserted 35 inches in th welder throat.

$a = 52$
$X_0 = 268 \times 10^{-6}$ ohm

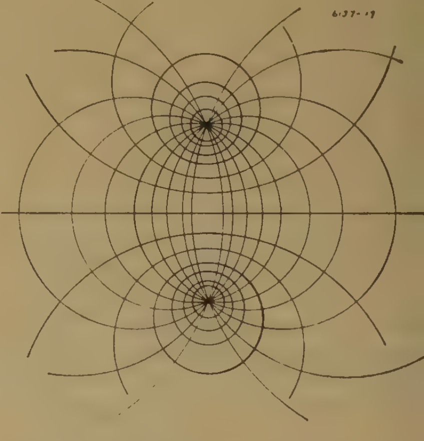


Figure 17. Graphical flow map of magnet field in air between two parallel conducto

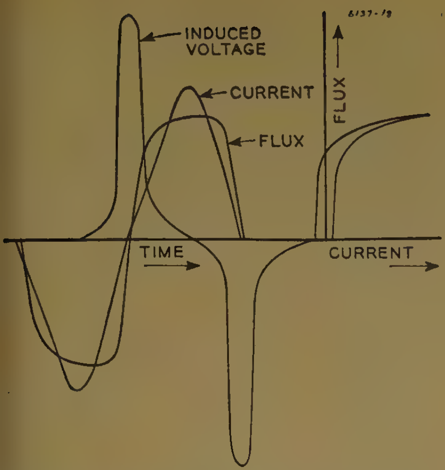


Figure 18. Schematic diagram of induced voltage, welding current, and flux in a saturated steel sheet

$R_o = 54 \times 10^{-6}$ ohm
$Z_o = 273 \times 10^{-6}$ ohm
$Z_p = 375/349$
$= 1.075$ ohms
$Z_s = 1.075/(52)^2 - 273 \times 10^{-6}$
$= 128 \times 10^{-6}$ ohm
$R_p = 26{,}700/(349)^2$
$= 0.219$ ohm
$R_s = 0.219/(52)^2 - 54 \times 10^{-6}$
$= 27 \times 10^{-6}$ ohm
$X_p = \sqrt{(1.075)^2 - (0.219)^2}$
$= 1.05$ ohms
$X_s = 1.05/(52)^2 - 268 \times 10^{-6}$
$= 121 \times 10^{-6}$ ohm
$I_t = 349(52)$
$= 18{,}100$ amperes
$V_{rs} = 18{,}100(27 \times 10^{-6})$
$= 0.488$ volt
$V_{xs} = 18{,}100(121 \times 10^{-6})$
$= 2.19$ volts

Appendix II. Applying Data to Specific Problems

Problem 1

Find the increase in secondary voltage necessary to maintain 18,000 amperes welding current in spot welding two sheets 0.120 inch thick and 36 inches wide when one inch and then 36 inches is inserted in the welder throat. Welder short-circuit constants are

$Z_o = 273 \times 10^{-6}$ ohm
$X_o = 268 \times 10^{-6}$ ohm
$R_o = 54 \times 10^{-6}$ ohm

From Figure 20

$I_t R_w = 1.26$ volts

From Figure 21

$V_{rs} = 0.86$ volt
Welder resistance drop $= I_t R_o$
$= (18{,}000)(54 \times 10^{-6})$
$= 0.97$ volt

Total resistive voltage drop $= V_r$
$= 3.09$ volts
Welder reactance drop $= I_t X_o$
$= (18{,}000)(268 \times 10^{-6})$
$= 4.82$ volts

From Figure 21

$V_{xs} = 2.87$ volts

Total reactive voltage drop $V_x = 7.69$ volts
Adding vectorially, $V_r + V_x = V_t$

$V_t = \sqrt{V_r^2 + V_x^2}$
$= 8.3$ volts at 36 inches
Voltage at start (one inch in throat) in which V_{rs} and V_{xs} are not effective
$= 5.32$ volts
Increase in voltage $= 8.3 - 5.3$
$= 3.0$ volts

Problem 2

The question is the same as in problem 1 except that the sheets are 18 inches wide, welding current is 10,000 amperes. The sheets are welded in the same welder. From Figure 20

$I_t R_w = 1.23$ volts

From Figure 21

$V_{rs} = 0.86$ volt

From this same chart this value must be corrected to

$K_1 \cdot V_{rs} = 0.65\ (0.86)$
$= 0.559$

because the sheet is only 18 inches wide. This value should be corrected further for change in base current, $K_3 = 1.05$.

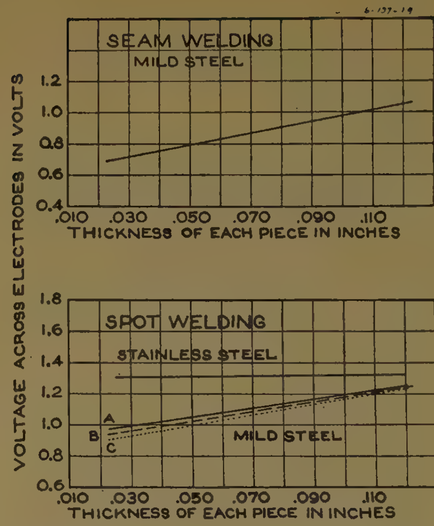
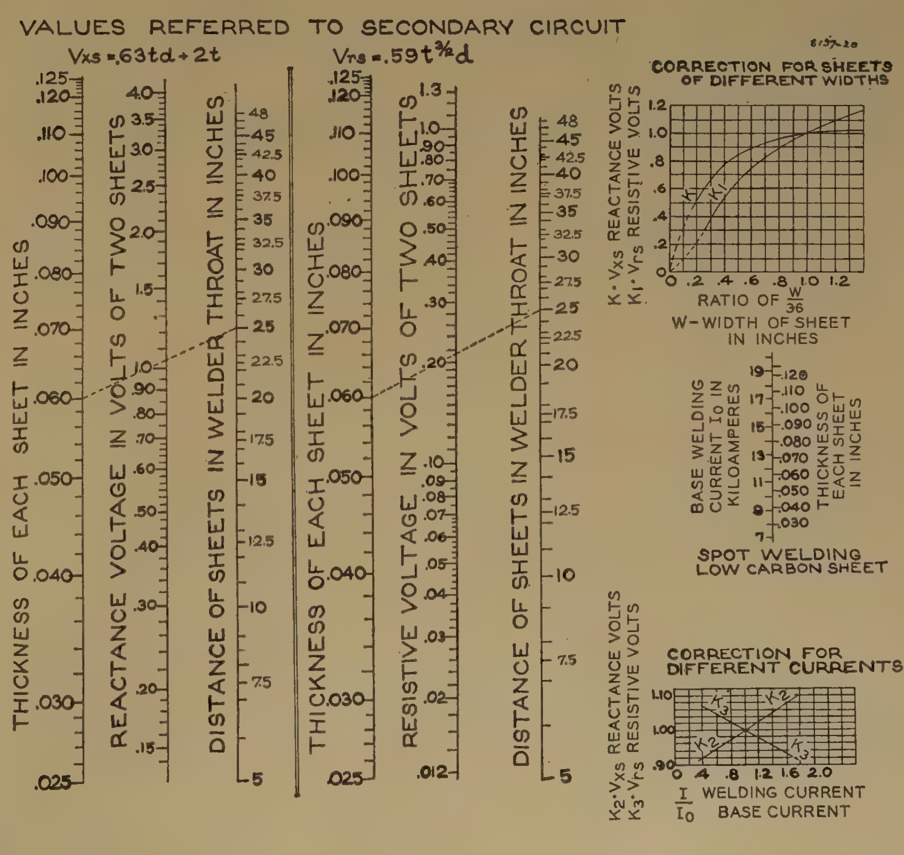


Figure 19. Resistive voltage drop across the electrodes caused by the welding resistance when welding small coupons

$K_1 \cdot K_3 \cdot V_{rs} = (0.65)(1.05)(0.86)$
$= 0.587$ volt
Welder resistance drop $= I_t R_o$
$= 10{,}000(54 \times 10^{-6})$
$= 0.54$ volt
Total resistive voltage drop $= V_r$
$= 2.36$ volts

Figure 20. Reactive and resistive voltage components caused by two thicknesses of low carbon steel 36 inches wide in welder throat



Welder reactance drop $=I_tX_0$
$=10{,}000\,(268\times10^{-8})$
$=2.68$ volts

From Figure 21

$K\cdot K_2\cdot V_{xs}=0.84(0.94)(2.87)$
$=2.27$ volts

Total reactive voltage drop $=4.95$ volts

$V_t=5.48$ volts at 36 inches

Voltage at start (1 inch in throat) in which V_{rs} and V_{xs} are not effective $=3.21$ volts

Increase in voltage $=5.48-3.21$
$=2.27$ volts

Appendix III. Determination of Flux and Flux Density From the Induced Electromotive Force or Voltage Wave

Basically

$$e=-N\frac{d\phi}{dt}10^{-8}$$

$$d\phi=\frac{-1}{N}e\cdot dt\frac{1}{10^{-8}}$$

$$\phi=\frac{-10^8}{N}\cdot\int_{t_1}^{t_2}e\cdot dt$$

where

N = number of turns on search coil
e = instantaneous voltage, volts
$\frac{d\phi}{dt}$ = rate of change of flux with time
ϕ = flux, maxwells or lines
t = time, seconds

The integral $\int_{t_1}^{t_2}e.dt$ represents the average voltage over the time interval considered. When the maximum value of flux is desired, as it was in this work, it is necessary to measure the area under the voltage wave e for one-half cycle by means of a planimeter, divide it by two, and substitute in the expression

$$\phi_M(\text{flux in lines or maxwells}) = \frac{\text{area of one-half of voltage wave in square inches}}{\text{number of turns on search coil}}\cdot\left(\frac{\text{volts}}{\text{inch}}\right)^*\cdot\left(\frac{\text{seconds}}{\text{inch}}\right)^*\cdot10^8$$

The maximum flux density B_m is found by dividing the flux by the area of steel that the coil encloses. This assumes that the flux is confined entirely to the steel or that the magnetic field in air is small compared to the magnetic field in the steel sheets.

$$B_m=\frac{\phi_m}{A}$$

where

A = area in square inches
B_m = maximum flux density in lines per square inch

* These are calibrations measured from the oscillogram of the voltage wave.

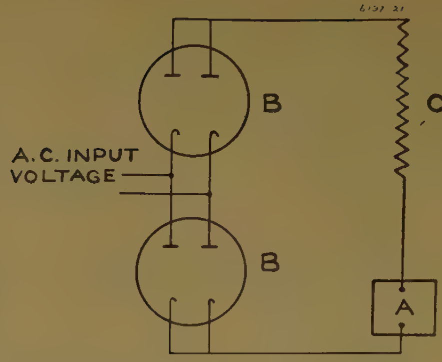


Figure 21. Rectifier unit for converting a-c input voltage to average direct voltage

A. D-c millivoltmeter, 1,000 millivolts, full scale, 5,000 ohms resistance
B. 6H6 tube
C. 100,000-ohm resistor

A direct reading meter can be made t evaluate $\int_{t_1}^{t_2}e.dt$ or the average voltage It was possible to measure this integral fo determination of ϕ_m by using a bridge-typ full-wave rectifier (Figure 22) and suitabl d-c meter. The rectified direct voltage i approximately 1/20 of the alternating volt age input. A search coil of ten turns o number 22 wire was used with the instru ment setup.

$$\phi_m=\frac{E_{avg}}{240N}\times10^8$$

where

$\frac{1}{240}$ = one-fourth of a cycle of a 60-cycl system
E_{avg} = average volts

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HIGHLIGHTS........

Summer Convention. With a registration of 800 for the first day, the summer convention gets under way in Detroit, Mich., as this issue goes to press. Full details of convention activities will appear in the August issue of *Electrical Engineering*.

Pacific Coast Convention. A full postwar program of eight sessions, plus entertainment, sports, and inspection trips, is planned for the Pacific Coast convention which will be held in Seattle, Wash., August 27–30, 1946. In anticipation of the continuing shortage of hotel accommodations, members desiring to attend should make hotel reservations immediately (*pages 329–30*). Complete details of the program will appear in the August issue of *Electrical Engineering*.

Control Council in Germany. The chief weakness of American policy in Germany "is the basic folly of the Potsdam Declaration and the unstable organization with which it is being implemented." This is the conclusion reached by AIEE Past President C. A. Powel who recently spent nine months in Germany as a member of the Allied Control Commission. As an alternative Mr. Powel proposes the adoption of the program devised by the engineering society presidents, a plan which would render Germany helpless to wage war but which would not impair her standard of living (*pages 306–09*).

Engineers and Management. In general the engineer in his relationship as an employee is well trained to handle the engineering aspects of his job and so he finds that his problems are typical employee problems whose solution lies in good management. It is when he becomes part of management himself, with the attendant need for general knowledge and understanding of human relations, that he becomes cognizant of the limitations of the average engineering education and its emphasis on technical training almost to the total exclusion of culture or the social sciences (*pages 322–5*).

V-2 Range Control. Information now is available concerning one of the more formidable enemy weapons, the German V-2 rocket, which has been undergoing tests by the United States Government at Las Cruces, N. Mex. Although no new basic knowledge is employed in the rocket which attained fame during the bombardment of England, the application is felt to be sufficiently novel to be of interest, particularly in the matter of range control (*pages 303–05*).

Employee Training Programs. With the prospect of almost 30 million persons returning to peacetime pursuits in the immediate future, industry is faced with a mass retraining problem as serious as that occasioned by a conversion to wartime production. Perhaps the example set during World War II by the government-financed Engineering Science and Management War Training Program will point the way to its solution (*pages 310–16*).

Wire Recording. Although there has been no widespread commercial exploitation of magnetic wire recording in the United States as yet, experiences with its use by the Armed Forces during World War II indicate a promising market for models adapted for commercial consumption. The simplicity of the recording procedure, the fact that the wire recording can be erased for reuse, and the length of playing time possible with a relatively small record all are factors which should find favor with the consumer in spite of various disadvantages existent in the recording in its present form (*pages 316–21*).

Conservation of Power. Improvements in power supply during the past two decades not only have decreased the unit cost of electricity to the consumer, but have accomplished much in the direction of conservation of fuel and other essential materials. However, in view of today's ever-increasing power requirements, further reduction in total fuel consumption becomes more and more difficult (*pages 326–8*).

Magnet Coils. With the development of silicone insulation, the distinction between insulation standards for magnet coils and those for rotating machines has become more pronounced. Based on this consideration, separate tests carried out on magnet coils reveal that a combination of fiber glass and silicone resin makes an important contribution to improving the thermal life of magnet coil insulation and seem to indicate that the end of the reliable life of such coils is determined by the failure of the silicone resins as bonds (*Transactions pages 412–16*).

Rectifier Installations. Rectifier installations may be classified either as *likely* or as *unlikely* to cause important induction problems, or as indeterminate. A working group of an AIEE subcommittee on electronic power conversion has prepared a report which presents methods by which proposed installations can be placed in one of these categories, thus saving considerable time and effort, either by omitting detailed studies and tests where they are not needed, or by taking the proper precautions when so indicated (*Transactions pages 417–36*).

Double-Reduction Motor. The modern double-reduction traction motor and gear unit incorporates qualities of smallness, compactness, and light weight which make it especially adaptable to a wide range of locomotive sizes where the weight saved in the traction motor may be used in the Diesel engine to provide a more powerful locomotive without increasing the weight. The speed of the double-reduction motor is approximately 67 per cent as much as that of the modern single-reduction motor which makes it unsuitable for heavier high-speed road locomotives (*Transactions pages 471–4*).

Diesel-Electric Drilling Equipment. With the application of the electrohydraulic governor, a war development, to Diesel-electric drilling equipment for the first time, a new degree of protection is attained through prevention of engine overloading. The governor also permits running the engines at the lowest speeds consistent with the desired rig motion and not at full rated speed at all times as in the past (*Transactions pages 447–53*).

Engineering Organizations. The necessity for an over-all engineering society to supplement specialized technical organizations by endeavoring to raise the professional standards of the profession as a whole is stressed in a current letter to the editor. Such a society, the writer believes, may be found in the National Society of Professional Engineers (*page 363*).

Directors' Report. Principal AIEE activities during the fiscal year ending April 30, 1946, are summarized in the 62d annual report of the board of directors. Also included are accountants' statements indicating the financial status of the Institute (*pages 331–50*).

Awards. National and District prize award winners were announced at the AIEE annual meeting in Detroit, June 26 (*page 351*).

News 329

ELECTRICAL ENGINEERING. Published monthly by the American Institute of Electrical Engineers; publication office 20th & Northampton Streets, Easton, Pa. Editorial and advertising offices, 33 West 39th Street, New York 18, N. Y. Subscription $12 per year plus extra postage charge to all countries to which the second-class postage rates does not apply; single copy $1.50. Entered as second-class matter at the Post Office, Easton, Pa., under the Act of Congress of March 3, 1879. Accepted for mailing at special postage rates provided for in Section 538, P. L. & R. Act of February 28, 1925. July 1946, Vol. 65, No. 7. Number of copies of this issue 31,000

HIGHLIGHTS........

August–September Issue. The double dateline on *Electrical Engineering* this month does not mean that an issue is being skipped, but is simply a device for rearranging the production schedule to insure early mailing of copies (*page 408*).

Summer Convention. Setting a new attendance record for recent years with a registration of 1,351, the 1946 AIEE summer convention was held in Detroit, Mich. An unusually heavy technical program marked this year's convention as well as a series of entertainment features highlighted by contests for the Mershon and Lee golfing trophies. This is the 61st annual summer convention of the Institute and the third time it has convened in Detroit (*pages 396–7*).

Sections. One of the outstanding features of the recent AIEE summer convention was a series of six conferences on Section activities arranged by the Sections committee. The conferences, whose popularity was attested by a very gratifying attendance, included an officers and delegates session which was addressed by both President Wickenden and President-Elect Housley (*pages 398–401*). Also included were three separate but parallel conferences which were held to permit discussion of common problems and exchange of experiences among Sections of similar size. The three sessions were conducted for the larger, intermediate, and smaller Sections (*pages 402–03*).

Professional Goals. "The essence of modernity is that progress no longer waits on genius; instead we have learned to put our faith in the organized efforts of ordinary men." In the presidential address delivered at the recent AIEE summer convention in Detroit, Mich., President Wickenden added his voice to the plea for national unity among engineers, one of the goals toward which Doctor Wickenden and his fellow AIEE officers have been working during the past year (*pages 365–9*).

Automatic Control Systems. Technical literature thus far has not given very adequate coverage to that class of servomechanism in which the components in the direct circuit are paralleled by components in a feed-back circuit. Treatment of a complex dynamic system is difficult on the basis of classical mathematics. However, an analysis is not only of value in giving a quantitatively approximate idea of the correct approach to a problem, but, if it includes experimental data on the performance of the major components, it can set up exact specifications for the design of the remaining circuits (*Transactions pages 521–9*). For an analysis of automatic control systems in general, the frequency response approach or the transient response approach may be used. Of the two, the frequency analysis provides information equivalent to the transient analysis, but with less calculation and more easily interpretable results (*Transactions pages 539–46*).

Physics' Half Century. Evolving over a period of 50 years from the periphery to the very center of the scientific circle, the United States today boasts more than three-fifths of the world's physics' activity—a far cry from an era when Benjamin Franklin and Joseph Henry were the only American physicists of note. Much of the current activity, of course, is the direct result of the recent war which stimulated scientific research in America but greatly curtailed it in Europe (*pages 378–83*).

Japanese Radar. Because "Our Emperor directed us to tell you everything," the Japanese people were unexpectedly co-operative in revealing to Allied officers what they could of Japan's technical activities during the war. What has been learned about their radar developments is reported in a 2-part article, beginning in this issue, by an ex-operations analyst who was assigned, after the fall of Japan, to study all phases of Japanese electronics and radar (*pages 370–7*).

Automatic Calculator. In 1642 the foundation for almost all modern mechanical calculating machines was laid when Blaise Pascal built the first mechanical adding machine incorporating the use of rotating wheels and providing for carry. One of the latest devices in this field, the automatic sequence controlled calculator developed by Harvard University and International Business Machines Corporation, is described in a 3-part article beginning in this issue (*pages 384–91*).

Quality Reports. The successful employment of quality reports in an increasing number of industrial concerns recommends an even more general usage although, in many cases, the installation of a quality control system will have to be preceded by the conversion of a hostile management. The reports themselves may be presented in any of several ways, the choice being determined by conditions in the plant in question (*pages 391–3*).

Radar Range. Most radar sets in use during the recent war functioned by transmitting a pulse and measuring the time necessary for the signal to reach the target and the echo to return. Under these conditions the most interesting problem which arises is the question of how much power must be transmitted so that the echo will be of sufficient strength to be detected satisfactorily by the radar receiver. In the case of microwave equipment, and the consequent line-of-sight transmission paths, the problem can be solved by geometric means (*Transactions pages 546–8*).

Ladder Networks. When using a uniform ladder network (in the 4-terminal sense) to simulate a smooth transmission line, it is agreed that under some conditions a mid-shunt termination gives the best results while with other terminal conditions the midseries arrangement is superior. With the understanding that the network retain its 4-terminal identity, can an intermediate termination be found which will give a uniformly good approximation for all conditions (*Transactions pages 530–6*)?

Dangerous Currents. Current and *not* voltage, in spite of prevailing misconceptions on that point, is the proper criterion of electric shock intensity. The effect of the shock on the victim will differ depending on the part of the body through which the current passes. The value of other factors, such as body and contact resistances or elapsed time between occurrence of the accident, rescue, and resuscitation usually can be estimated only roughly (*Transactions pages 579–85*).

Preserving Pole Woods. As a result of the backlog of pole requirements built up during the war and of today's labor shortages, it has become difficult to obtain a sufficient supply of the more extensively used pole woods. In an effort to solve the problem the American Standards Association has issued specifications covering various previously unrecognized conifers which may be given preservative treatments and used as substitutes (*Transactions pages 549–54*).

News 394

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HIGHLIGHTS........

Pacific Coast Meeting. With an estimated attendance of more than 500, the Pacific Coast meeting was held this year in Seattle, Wash., August 27–30. Full details of the meeting will be reported in *Electrical Engineering* for November.

Japanese Radar. The insistence of Japanese military leaders on keeping army and navy research, development, production, and operation entirely separate, as well as Japan's failure to organize the full scientific power of the nation for war research, probably was largely responsible for her inability to develop effective radar equipment during World War II. This conclusion is reached in Part II of an article based upon an official report prepared by the United States Army (*pages 455–63*).

Lamme Medal Award. The Lamme Medal for 1945 was awarded to David C. Prince (F '26) for his work with high voltage switching equipment and electronic converters. The three addresses delivered during the medal presentation ceremony include some pertinent remarks concerning the donor, Benjamin G. Lamme; an account of the medalist's career; and Doctor Prince's acceptance address in which he acknowledges the inspiration afforded by the inscription on the Lamme Medal, "The engineer views hopefully the hitherto unattainable" (*pages 435–41*).

Statistical Methods. The 11th article in the statistical series discusses the statistical interpretation of limited experimental tests. These "tests of significance," as they generally are called, have proved useful disciplines in experimental development although they often are misused through superficial generalization and enthusiasm (*pages 466–8*).

Unionization of Engineers. Although of the personal opinion that unionization of engineers is an unnatural development, an AIEE Student Member warns the profession that the labor union idea is becoming increasingly widespread among young engineering graduates. He believes that one solution might lie in a more active interest on the part of established engineering societies in the employment difficulties of their members (*pages 445–8*).

Westinghouse Centennial. Since the birth of George Westinghouse 100 years ago the era of rugged individualism in which he was a leader has given way to a period of increasing dependence on planned economy and government control. Technologically and scientifically the intervening years have witnessed much progress, but they also have witnessed a serious undermining of those elements so essential to the continuance of a properly functioning society; namely, individual responsibility, good will, and leadership. If Westinghouse were alive today he surely would agree that progress is attained primarily through individual liberty (*pages 442–5*).

Automatic Calculator. Part II of a 3-part article describes the functions performed by the multiplication and division registers and the functional units of the automatic sequence controlled calculator presented to Harvard University (*pages 449–54*).

Bikini Report. Were the atomic bomb tests at Bikini worth while? What are the general indications for the future? A member of the Institute who witnessed the tests relates his observations, which coincide, in general, with official reports on the subject (*pages 463–5*).

UN Electrical Facilities. Reputed to be the largest continuously louvered ceilings in existence, the ceilings of the two main council chambers of the Lake Success quarters of the United Nations each consume a total of 24 kilowatts for illumination. Each chamber has a general illumination of 58 foot-candles at the center of the room which tapers to 35 foot-candles at the edges (*page 484*).

Electrical Engineering Exposition. An electrical engineering exposition originally planned for 1941 but postponed because of the war will be held in New York concurrently with the AIEE 1947 winter meeting. The exposition will feature the newest developments in electric equipment for the generation, transmission, distribution, and utilization of electric energy (*page 485*).

News 469

Mica Capacitors. Current design of carrier telephone and other communication systems imposes special demands on capacitors used in frequency-sensitive systems. To meet these exacting requirements for capacitance tolerance at the time of manufacture and for subsequent permissible change in capacitance, mica capacitors having sprayed silver electrodes bonded to the mica laminations have been developed to replace former designs (*Transactions pages 670–4*).

Aircraft Engine Synchronism. With increasing emphasis being placed on multiengine airplanes for commercial and military use, it has become more and more important to devise a satisfactory method for synchronizing aircraft engines in order to attain the maximum in comfort and flight efficiency, and visual synchronism seems to meet the requirements. This method employs an instrument which indicates the degree of synchronism between two engines by measuring the frequency difference between the outputs of the generators connected to each engine (*Transactions pages 650–3*).

Current Ratings for Wire and Cable. There is no accepted standard method for determining the short-time current rating of wires. However, based upon a comparison of the results of the three methods generally used by engineers—thermocouple, oscillograph, or voltmeter-ammeter—a current-time-temperature test is recommended which utilizes a voltmeter-ammeter for times down to five seconds and an oscillograph for shorter times (*Transactions pages 644–8*).

Servomechanism Analysis. The mechanical transients analyzer provides a fast and simple method of obtaining complete solutions of the performance equations of servomechanisms under any type of operating condition. Solutions based on data obtained from a general study of variable-voltage angular-position servomechanisms include system frequencies and per cent damping per cycle that are applicable to any system disturbance (*Transactions pages 636–9*).

Ignitrons. Because the auxiliaries required for excitation and temperature control are an important part of the rectifier unit, the successful design of this unit must allow for the characteristics, limitations, and requirements of the electron tube. Among the factors which must be included in application data for mercury pool tubes or tanks are their excitation, control, and cooling characteristics (*Transactions pages 632–5*).

Politics. AIEE member Victor Wichum is a candidate for Congress from the 10th Congressional District, Kings County (*page 492*).

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HIGHLIGHTS........

Great Lakes Meeting. As this issue goes to press, reports from Indianapolis, Ind., where this year's Great Lakes District meeting is being held, October 9–11, indicate registration of well over 250. Full details of the meeting will appear in the December issue.

Pacific Coast Meeting. Offering a well-balanced program of nine sessions, including a technical session on atomic power, the recent AIEE meeting at Seattle, Wash., attracted an attendance of 562, a record for Pacific Coast meetings. Among the highlights of the meeting were an address on the "Development of Atomic Power" by Bruce R. Prentice, and a session devoted to a discussion of the organization of the engineering profession (*pages 529–31*).

John Fritz Medal Awarded. Announcement has been made of the award of the John Fritz Medal for 1947 to Doctor Lewis Warrington Chubb (F '21) director of research in the research laboratories of the Westinghouse Electric Corporation, East Pittsburgh, Pa. Doctor Chubb was selected for his "pioneering genius and notable achievements during a long career devoted to the scientific advancement of the production and utilization of electric energy." The December issue of *Electrical Engineering* will have a further account of Doctor Chubb's career.

National Electronics Conference. Reports from the National Electonics Conference, meeting in Chicago, October 3 to 5, give the initial day's registration as exceeding 1,800. A full report of the 19 technical sessions of the conference will appear in the December issue.

Reactive Power Sign. Reader comment is invited on a proposition to change the sign of inductive reactive power from an arbitrarily accepted negative value to positive in order to provide a standard which will express the term "reactive power" as it generally is employed in engineering practice. Justification for the change is based on a report prepared by a subcommittee of the AIEE Standards committee (*pages 512–16*).

Industrial Hazard. The usefulness of X and gamma radiations in industry is partly offset by the hazard to workers engendered by the rays' "delayed-action" destructive effect on living tissue. Because it would be uneconomical if not impossible to make all industrial equipment employing the radiations absolutely foolproof, it is important that employees be provided with a set of supplementary rules to be obeyed for adequate protection (*pages 499–507*).

Science, Strength, Stability. Government by the people is really government by that portion of the people which takes the trouble to participate actively in the formation of public opinion. The engineer and the scientist, because of the specialized positions they, as professional men, hold in society, play a leading part in the development of that public opinion and so must accept their share of the responsibility for the enlightened world opinion which is a necessary condition for world peace (*pages 508–12*).

Development of the Saint Lawrence. Construction of a navigation channel 800 feet long and the development of 2,200,000 horsepower of hydroelectric capacity are two features of a proposed plan for the hydroelectric development of the Saint Lawrence River. The project, when completed, will provide power and navigation benefits for both the United States and Canada under terms of an agreement between the two nations for co-operation in its construction and maintenance (*pages 495–8*).

Human Progress. Although the increasing destructiveness of recent wars would seem to bode ill for the future of civilization, a survey of the progress of mankind since the era of Peking man still leaves room for optimism—that is, if man can learn to profit by the lessons of the past (*pages 521–2*).

Relays for Telegraphy. From an initial trial installation with a 3-link system from New York to Philadelphia, much progress has been made in the past year in the development of a radio relay system for domestic telegraphy. By materially reducing circuit interruptions resulting from storms, falling trees, and electrical disturbances, such a system is expected to improve the quality, dependability, and speed of telegraph service provided to the public (*pages 516–20*).

Automatic Calculator. The third and concluding part of an article describing the automatic sequence controlled calculator discusses the preparation and planning of the sequence control tapes. These tapes are employed when a large number of values is to be used by the calculator in a prescribed order. The values may be supplied to the machine via a perforated paper tape and one of the interpolator mechanisms to each of which three sequence codes have been assigned (*pages 522–8*).

Philosophy of Relaying. From a strictly objective point of view, the electric power system exists to render service to the customer better and more economically than he could be served by other methods. In spite of all precautions to assure excellence of service, however, the system still remains subject to unavoidable defects which result in faults that may damage equipment and interrupt service. Although relay protection cannot prevent these faults from occurring, it is a means of minimizing the effects of such faults when they do occur (*Transactions pages 735–41*).

Thermistors. The thermistor (or thermally sensitive resistor), a new circuit element made of solid semiconducting material, has characteristics of versatility and durability which make it especially adaptable for such applications as sensitive thermometers and temperature control elements, gas pressure gauges, and contactless time delay devices. Previously found impractical for widespread use because of a wide variation in the properties of units seemingly produced by the same process, the thermistor's present possibilities are the result of almost a century of research to overcome the device's imperfections (*Transactions pages 711–25*).

Military Communications. One of the major differences between the operation of a military communication system in the field and commercial operations is the method by which the requirements are limited with respect to the available supply of facilities. To develop a theater-wide system of communications with the complex operating features involved in a military setup, it soon was realized that staff responsibility without technical administrative authority over the operating personnel of the various subordinate command was entirely unsatisfactory (*Transactions pages 757–61*).

News..................529

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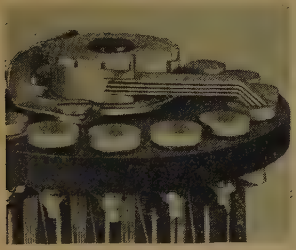
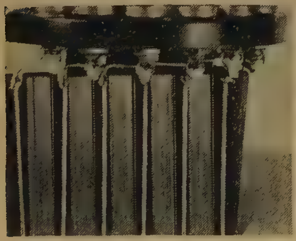
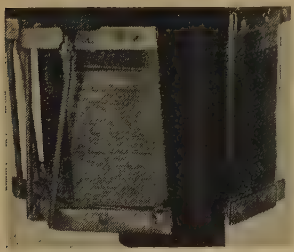


Quadruple-leaf, phosphor-bronze switches with wiping contact. Contact surfaces ground flat. Contact resistance only 0.002 ohm. Cam-type positive detent mechanism. Entire unit shielded behind panel.



Aryton-Perry windings of the 1-, 10-, and 100-ohm decades on bakelite cards.



Unifilar windings of the 1,000- and 10,000-ohm decades on mica cards.

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HIGHLIGHTS........

New Publication Policy. To meet the expressed desires of the membership for a broader and more effective publications service, important changes in AIEE publication policy and procedure will become effective January 1, 1947, in accordance with recommendations made by the publication committee and approved by the board of directors, October 26, 1946 (*pages 576–8*).

Great Lakes District Meeting. In spite of a last minute change in location from Fort Wayne, Ind., to Indianapolis, Ind., as a result of the hotel situation in the former city, the seventh annual meeting of the AIEE Great Lakes District, October 9–11, was carried to a successful conclusion. Featured at the meeting was a general conference on Institute activities and organization of the engineering profession, under the chairmanship of AIEE Vice-President T. G. LeClair of the Great Lakes District, which was addressed by AIEE President J. Elmer Housley (*pages 578–80*). The majority of the papers presented at the meeting were conference or District papers only. Their subject matter included electric machinery, electronics, communications, and basic sciences (*pages 565–9*).

National Electronics Conference. Approximately 2,100 persons attended the second National Electronics Conference which was held in Chicago, Ill., October 3–5. This conference was organized in 1944 as a national forum on electronics developments and their applications to keep abreast of recent rapid growth in that field. The AIEE Chicago Section is one of its sponsors (*page 590*). As an indication of the nature of subjects discussed at the conference, digests of the papers are presented in this issue (*pages 569–74*).

Science and Politics. It is acknowledged generally that the union of science and government during World War II was extremely successful in producing the scientific developments which largely determined the winning of the war. Does the success of this partnership in meeting the abnormal demands of war argue its continuation as insurance of an adequately financed peacetime program? One argument in its favor points out that, inasmuch as the public eventually benefits from scientific research, it has a direct interest in supporting it. This is a debatable subject of immediate and far-reaching importance (*pages 554–6*).

Train Communications. Tests of inductive and radio carrier communications for trains conducted during the past few years under actual operating conditions indicate that such systems, although not entirely satisfactory, are valuable as time-saving devices. The tests were carried out on major United States railroads in co-operation with radio equipment manufacturers, the Association of American Railroads, and the Federal Communications Commission (*pages 547–53*).

Organization of the Engineering Profession. Because there are two aspects to be considered in the organization of the engineering profession—the technical side and the professional side—it would seem that Plan *B* of the four plans submitted by the professional activities subcommittee of the AIEE committee on planning and co-ordination presents the best course to follow. This plan leaves the AIEE and other technical societies free to continue their technical activities and calls for a professional society to carry on all activities of a nontechnical nature (*pages 563–4*).

Slow-Acting Relays. Slow-acting relays have many uses, but their widest application is in automatic telephone systems where they are an essential element. Although little has been said on the subject in technical literature, the advantages of the slow-acting relay, which include small space requirements, low cost, and low energy requirements, far outweigh their only major disadvantage, a question of mounting, and even this has been overcome to a large extent by modern design (*pages 557–63*).

Fault-Current Measurement. A fault-current measuring device must be rapid and positive in action, must not interfere with the electrical characteristics of protective apparatus in the same circuit, and must be prepared to operate automatically. One device which meets these requirements is a magnetic link measuring device which utilizes a surge-crest ammeter and a demagnetizing coil to read the magnetization of the links and to demagnetize them after energization (*Transactions pages 839–43*).

News....................575

Radio Relay System. Although designed primarily for military use, an 8-channel microwave relay system which utilizes radio frequencies approaching 5,000 megacycles provides such complete freedom from static and most man-made interference that its basic design principles could be employed profitably in commercial applications (*Transactions pages 798–806*).

Aeronautical Research. The aeronautical research laboratories of the United States are of interest, not only for the part they play in aeronautical developments, but for the variety of electric devices they employ, and the surprising magnitude of their electric power loads. For example, the Ames Aeronautical Laboratory at Moffett Field, Calif., one of three laboratories of the National Advisory Committee for Aeronautics, will have more than 150,000 connected horsepower when wind tunnels now under construction have been completed (*Transactions pages 833–9*).

Teaching Electricity. To comprehend fully the significance and limitations of each of the relations, formulas, and rules which make up the body of a science, it is necessary to understand the basis upon which each was obtained. Thus, a presentation of most of the fundamental relations in electricity and magnetism in general form, with the basis of each relation indicated clearly, should provide the beginning student with the means for acquiring a fundamental understanding of electricity and magnetism, and forms his basic training in engineering analysis (*Transactions pages 828–33*).

Corrections. Attention is called to the following misprints occurring in recent issues of *Electrical Engineering*. In the paper "The Influence of the Concentration and Mobility of Ions on Dielectric Loss of Insulating Oils" by Bun Po Kang, published in the Transactions section of the July 1946 issue, on page 403, column one, fourth line from the bottom, "within a limited rage of temperature" should be corrected to read "with a limited range of temperature," and equation 3 on page 404 which appears as "$W=A-B/T$" should read "Log $W=A-B/T$." On page 458 of the same issue, also in the Transactions section, an error appears in the paper "Operating Experience With Distance Ground Relays" by W. A. Wolfe. The second term of equation 2 should read $I_{0GF}\frac{Z_{0GH}-Z_{1GH}}{Z_{1GH}}$ instead of $I_{0GF}\frac{Z_{0GH}+Z_{1GH}}{Z_{1GH}}$ as printed. In the paper "Application Ratings of Indoor Power Circuit Breakers" by O. B. Vikoren, published in the November issue, the 14th line, second column, page 270 of the Transactions section, which reads "in the 400-ampere and 440-ampere tests" should be corrected to "in the 400-ampere and 550-ampere tests."

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DECEMBER

1946

TRANSACTIONS SECTION

(Follows EE page 600; a preprint of pages 779–860 of the 1946 volume. Any discussion of these papers will appear in the December 1946 "Supplement to Electrical Engineering—Transactions Section" and in the 1946 volume.)

Statements and opinions given in articles and papers appearing in *Electrical Engineering* are the expressions of contributors, for which the Institute assumes no responsibility. ¶ Correspondence is invited on all controversial matters.

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NUMBER 12


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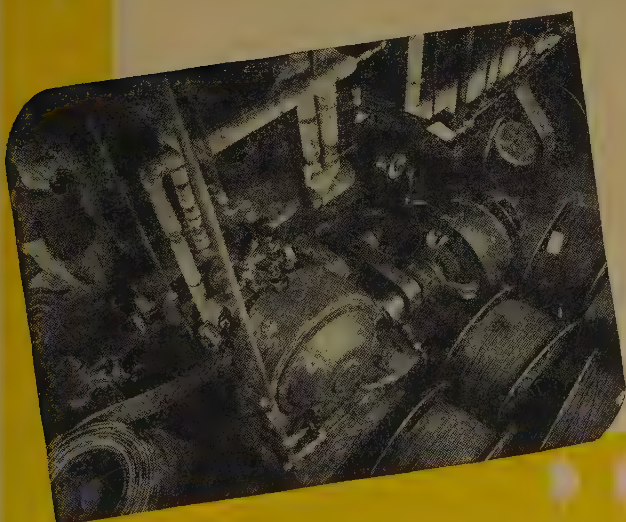

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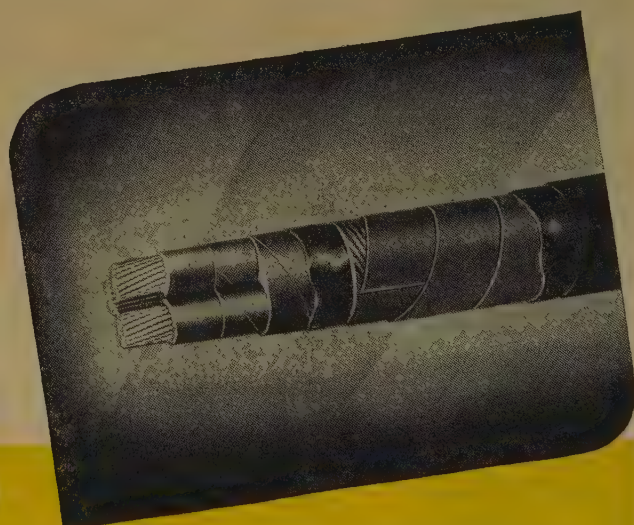

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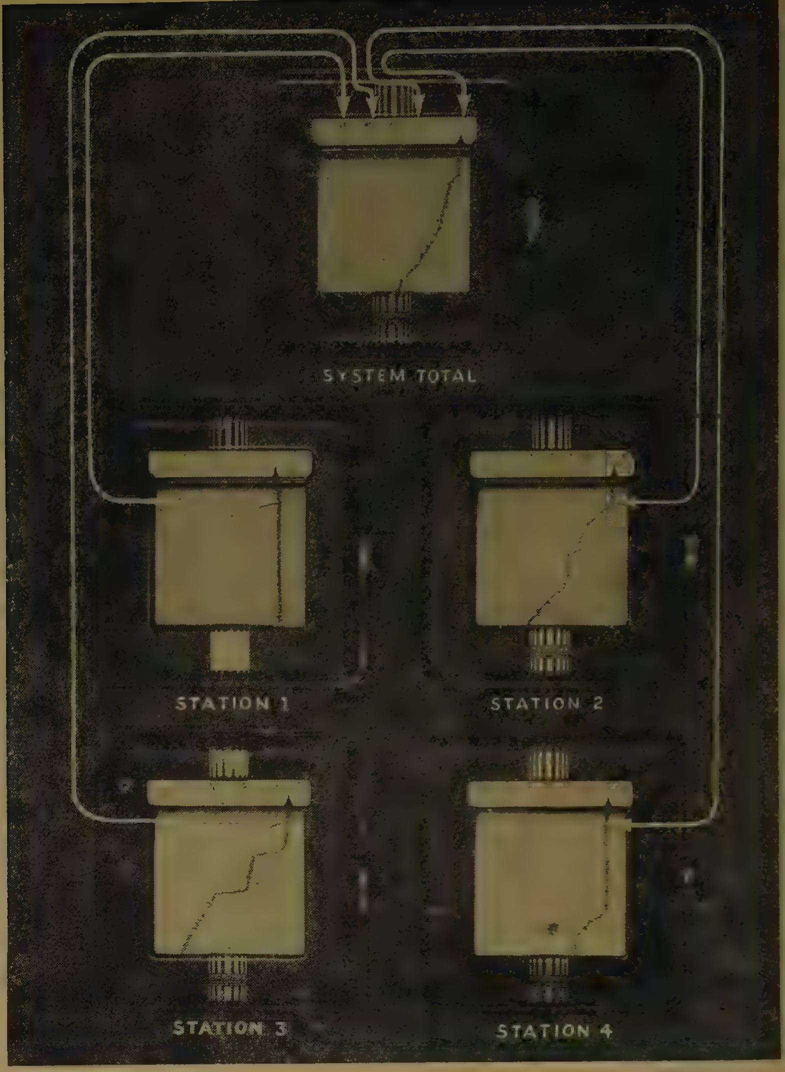

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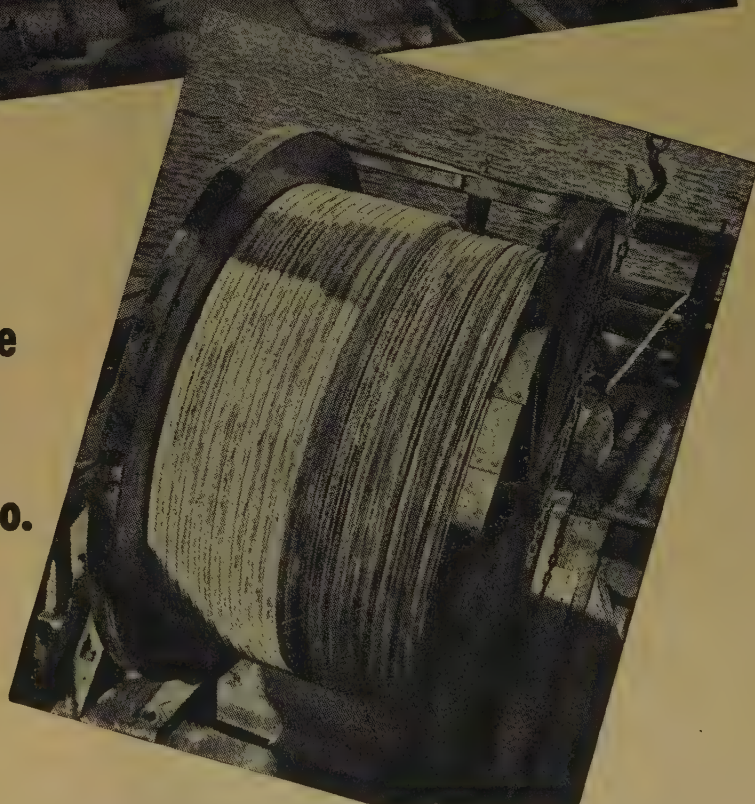

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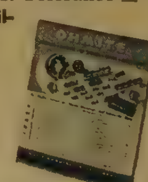

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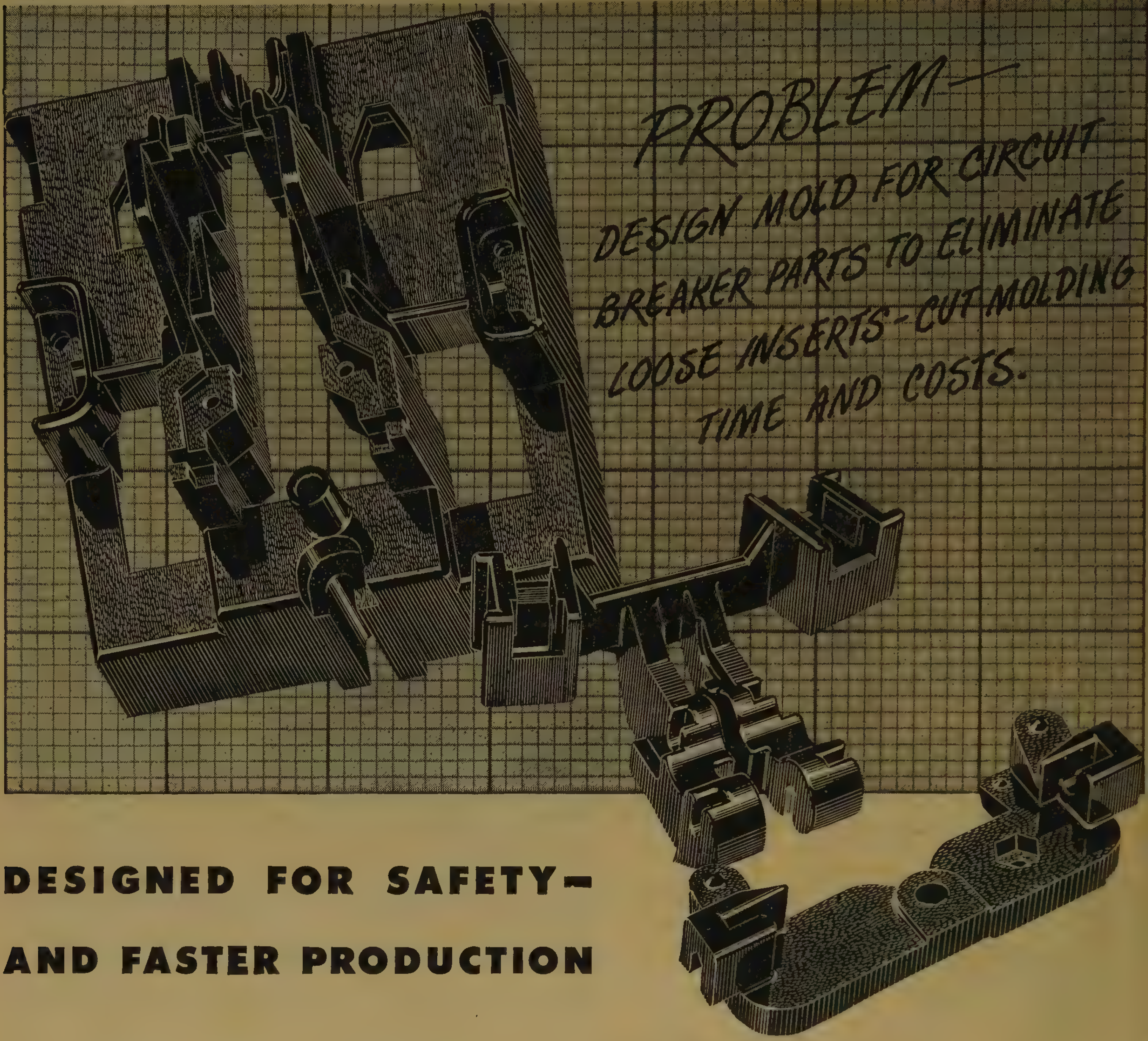

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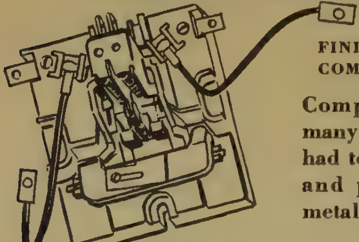

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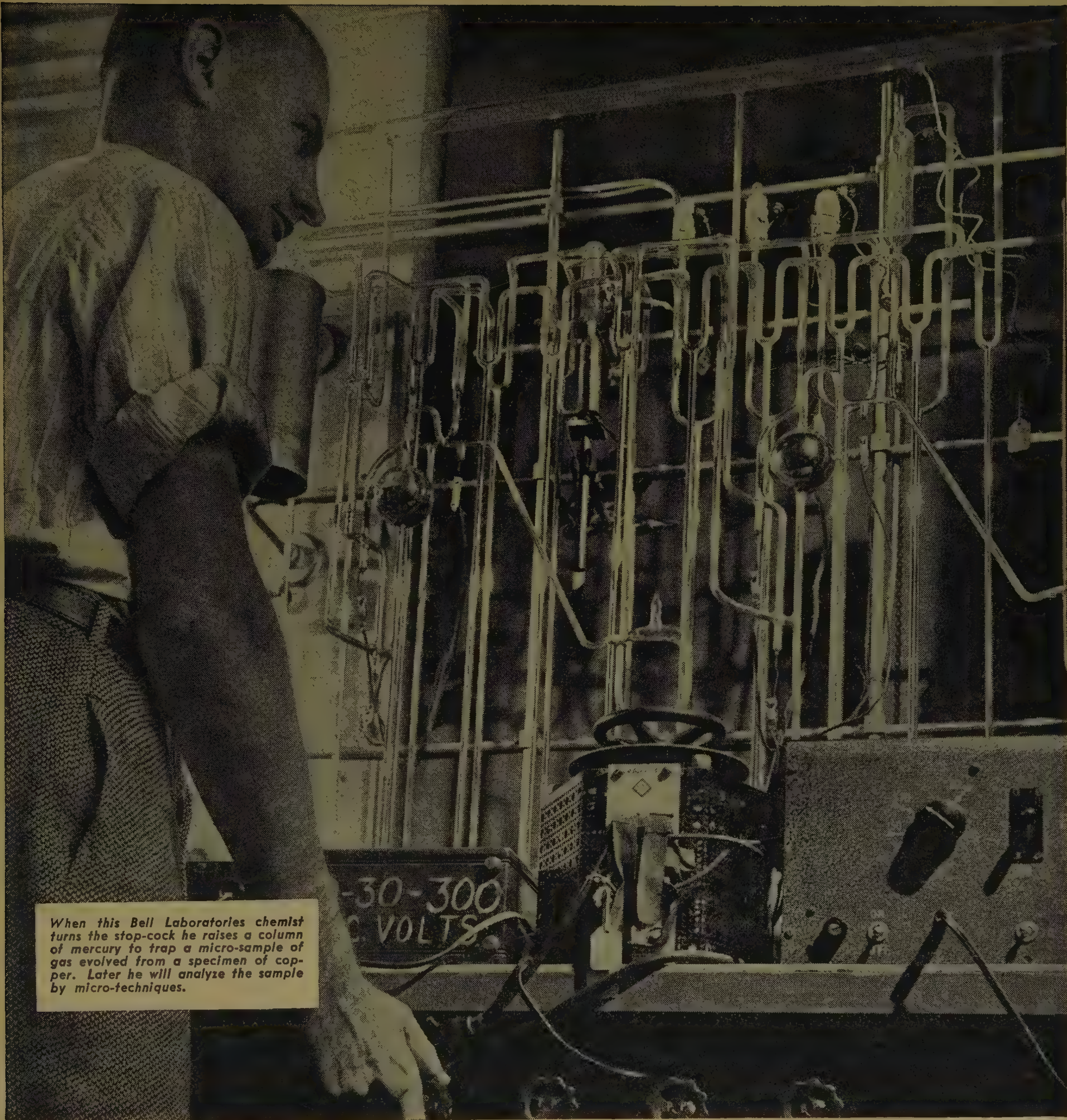

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EXPLORING, INVENTING, DEVISING AND PERFECTING FOR CONTINUED IMPROVEMENTS AND ECONOMIES IN TELEPHONE SERVICE

SERVING THROUGH SCIENCE

FARM OR HOSPITAL

Be Sure to Ask this Question

SMALLEST DIAMETER, LIGHTEST WEIGHT NATURAL RUBBER INSULATED WIRE

- **SAFER** because of perfectly centered conductors.
- **SAFER** because 10 layers of pure rubber insulation guard against current leakage.
- **SAFER** because of special fibrous, flame-resistant cover.

Smaller size permits more circuits in conduit.

U.S. Laytex Reg. U.S. Trade Mark **CONSTRUCTION**

RU wiring with RU?

Adequate wiring—for new structures or old—is assured if the specifications read "RU-Laytex".

Electrical wholesalers, architects, engineers and contractors, familiar with this famous product of U. S. Rubber engineering, know from experience that here is the wire for modern electrification.

For RU-Laytex, with its unique insulation of purified natural rubber, is today's lightest weight, smallest diameter, rubber insulated building wire.

Both in physical and electrical properties RU-Laytex leads the field. It prevents current leakage, has greater resistance to climatic deterioration, is easier to install, permits more circuits per conduit.

So keep that question in mind: "RU wiring with RU?"
The right answer means greater satisfaction for all concerned.

(RU type is approved by the National Electrical Code as a general all-purpose wire.)

UNITED STATES RUBBER COMPANY TRADE MARK

UNITED STATES RUBBER COMPANY

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We build towers
of EVERY TYPE
Need transmission towers? Substations? Radio towers? Floodlight towers? Even a power windmill or memorial tower? Whatever your tower requirements, American Bridge Company has unequalled facilities and engineering experience to do the job right. It will pay you to send us your inquiries.
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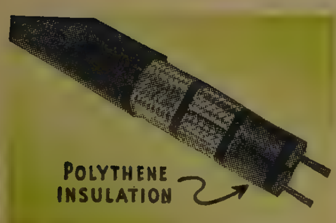
Surer even than a pilot's eyes

...the high-freauency cable that makes **RADAR** *possible*

is insulated with

Du Pont POLYTHENE



CUT-BACK SECTION of a modern twin coaxial cable for use in direction-finding equipment at ultra-high frequencies (400 to 10,000 megacycles). JAN-C-17 designation RG-23/U; made by Anaconda Wire & Cable Co. The insulation, about 140-150 mils thick, is Du Pont polythene.

Radar and associated electronic developments, born in stress of war, have important peacetime uses immediately ahead. Ships and planes which, without radar, might be hopelessly lost in foul weather or fog, may now soon be located ... come safely to harbor or airport ... thanks to one or several of the four main types of radar.

In this important work Du Pont polythene plays an indispensable part. For radar depends on ultra-high frequencies—anywhere from 2 to 10,000 megacycles! The cables that carry these currents must have an insulating material of low electrical losses—which at the same time is flexible at low temperatures. *Du Pont polythene meets both these requirements.*

Other properties of Du Pont polythene: outstanding chemical inertness, light weight (specific gravity 0.92), low moisture permeability, flexibility in thin sections, rigidity in thick. For complete data sheet, write to E. I. du Pont de Nemours & Co. (Inc.), Plastics Dept., Room 1612, Arlington, N. J.

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Plastics

Du Pont manufactures polythene molding powder.

Commercial extruders convert polythene into the forms of

SHEETS **TUBES** **RODS** **FILAMENTS**

The New Way to *Insulate and Protect* harnesses and leads

WRAP WITH Fibron PLASTIC TAPE

Transparent Fibron Plastic Tape permits ready inspection. Plastic gaskets also available.

Whenever equipment wiring must withstand severe operating conditions, attacks from moisture, oils, and chemical fumes as well as mechanical abuse, wrap harnesses and leads with protective Fibron Plastic Tape. This black or transparent tape is tough, remains flexible at low temperatures, and has good insulating properties.

Extremely elastic, Fibron Plastic Tape wraps smoothly and evenly over irregular surfaces and will not bulk in sharp corners.

Can be Fused into Homogeneous Structure

Ends can be instantly bonded by heat sealing, or the completely wrapped harnesses infra-red or oven baked to form a solid mass of insulation.

Note the outstanding characteristics of Fibron Plastic Tapes as listed below. Plan to test these products now. Generous samples will be gladly sent on request.

Outstanding Properties of Fibron Plastic Tapes

	#1 Black	#3 Transparent
Thicknesses	.008″ .012″ .020″	.020″
Widths	½″ to 3″	½″ to 3″
Dielectric Strength (.012″ tape)	1000 VPM	750 VPM
Tensile Strength, lb. per sq. in. (dumbell specimen tested)	1600	1600
Elongation	250% to 400%	350%
Brittleness Temperature	−41 deg. C.	−38 deg. C.
Bonding Temperature (depending on method used)	130°-150°C.	135°-155°C.
Specific Gravity	1.25	1.22

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Elevation 13,500 feet...

AND TOUGH GOING TO GET THERE . . . WOULD BE A "NATURAL" FOR TOWERS OF ALCOA ALUMINUM

The lighter weight of Alcoa Aluminum structural shapes greatly simplifies the transportation of materials to sites like this. It makes erection easier, too—facts well worth remembering when you're designing towers for out-of-the-way places.

Even if you don't have a right-of-way over this kind of country, there's another way Alcoa Aluminum can simplify your operation.

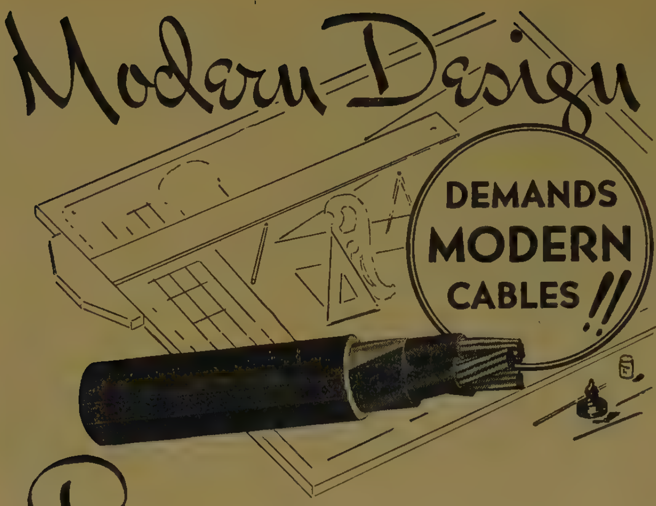
Aluminum is highly resistant to corrosive attack; it normally needs no protective painting. Use Alcoa Aluminum structural shapes, therefore, in the upper portions of your towers and eliminate the need for costly and hazardous painting.

Designers and operating men interested in this idea may obtain full data on Alcoa Aluminum structural shapes through the nearby Alcoa office. Or write ALUMINUM COMPANY OF AMERICA, 2149 Gulf Building, Pittsburgh 19, Pennsylvania.

Aluminum normally needs no protective painting.

ALCOA FIRST IN ALUMINUM

IN EVERY COMMERCIAL FORM



PLANS for today's airport, building, and highway construction are thoroughly modern. They call for underground power lines, and with an eye on increased performance at decreased expense specify ANHYDREX* underground cables.

Simplex - ANHYDREX underground cables may be installed in ducts or buried directly in the soil without any outer coverings.

A tough rubber jacket provides protection against the injurious effects of abrasion and acidity, and eliminates the sheath currents and electrolysis common to metallic-covered cables.

The ANHYDREX insulation is the most moisture-resistant rubber compound known to the wire and cable industry and even constant immersion in water will not impair either its physical or electrical properties.

It never pays in savings or performance to be behind the times, but it will to specify ANHYDREX cables for your next construction job.

*Trade Mark Reg. U. S. Patent Office.

Simplex Wire & Cable Co.
79 Sidney Street, Cambridge, Mass.

At right: Aerovox Series 20 Hyvol impregnated and filled capacitors in ratings up to 50,000 v. Below: Aerovox Series 26 stack-mounting Hyvol impregnated and filled capacitors in ratings up to 150,000 v.

● For those extra-severe-service applications of the largest capacitors, as well as others, Aerovox units have that extra stamina that makes them last. Decades of specialization provide an experience background second to none in solving all kinds of capacitor problems; unexcelled production facilities assure QUALITY as well as quantity.

Aerovox capacitors are liberally engineered for their individual applications. Special multi-layer capacitor tissues . . . long-life, non-inflammable Hyvol impregnant and fill . . . constant filtration as regular production routine . . . thorough evacuation and impregnation . . . positive hermetic sealing—these facts of Aerovox craftsmanship spell long, trouble-free service.

Aerovox capacitors in daily use speak for themselves. No finer capacitors are built. Aerovox engineers stand ready to meet your most severe requirements.

● *Submit your capacitor problem. The tougher the better! Write for literature.*

AEROVOX capacitors FOR RADIO-ELECTRONIC AND INDUSTRIAL APPLICATIONS

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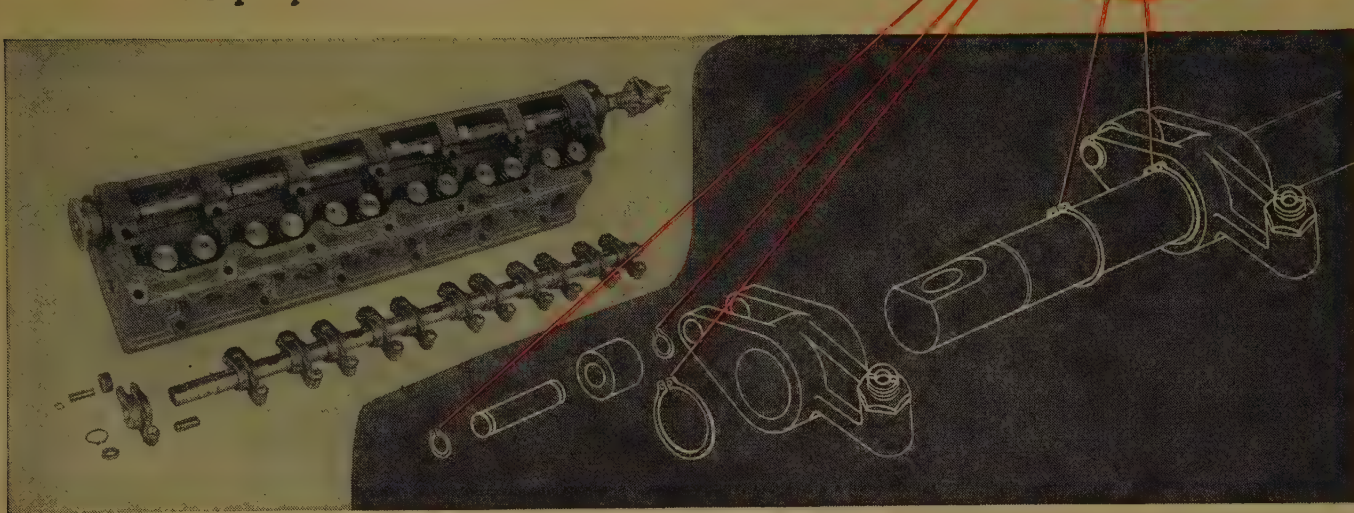
Twin Coach uses Truarc Rings on new bus engine—reduces weight ratio to 1 hp. per 4.7 lbs.

48 WALDES TRUARC RETAINING RINGS ON NEW ROCKER ARM

- **Slash Labor Costs**
- **Cut Production Time**
- **Guarantee Accuracy**
- **Simplify Maintenance**



"WE'VE FOUND TRUARC A NATURAL FOR AUTOMOTIVE APPLICATIONS!" reports Twin Coach Company, of Kent, Ohio. Their new Fageol Twin Coach engine uses a total of 52 Waldes Truarc Retaining Rings—48 on rocker arm, 4 on water pump and oil pump drive—to develop the remarkable efficiency of 1 hp. per 4.7 lbs. (the average pre-war gasoline bus engine produced 1 hp. for each 9 lbs., diesel engine 1 hp. for each 10 lbs.)

Automotive designers are specifying Truarc rings in transmissions, clutches, brakes, steering mechanism, and other vital assemblies. Production and maintenance men find that Truarc cuts costs sharply, is a superior solution to fastening problems because of its never-failing grip, its patented design assuring constant circularity. Send us your drawings; Waldes Truarc engineers will be glad to show how Truarc can help you.

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WALDES TRUARC RETAINING RINGS

TRADE MARK U. S. PAT. RE. 18,144

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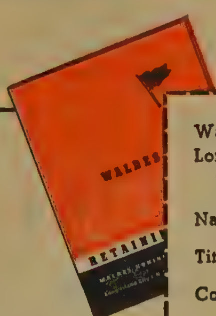

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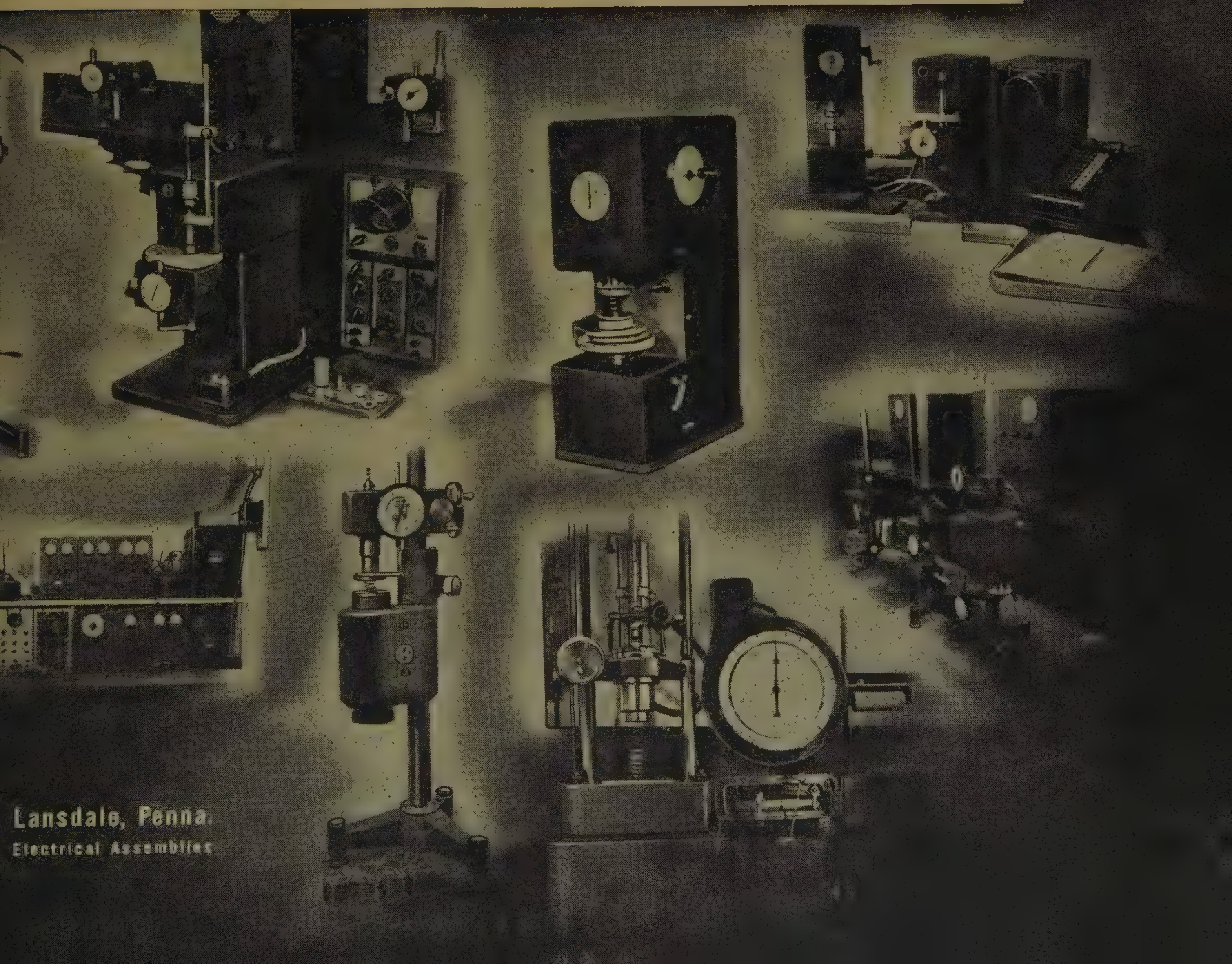
All this to test Springs?

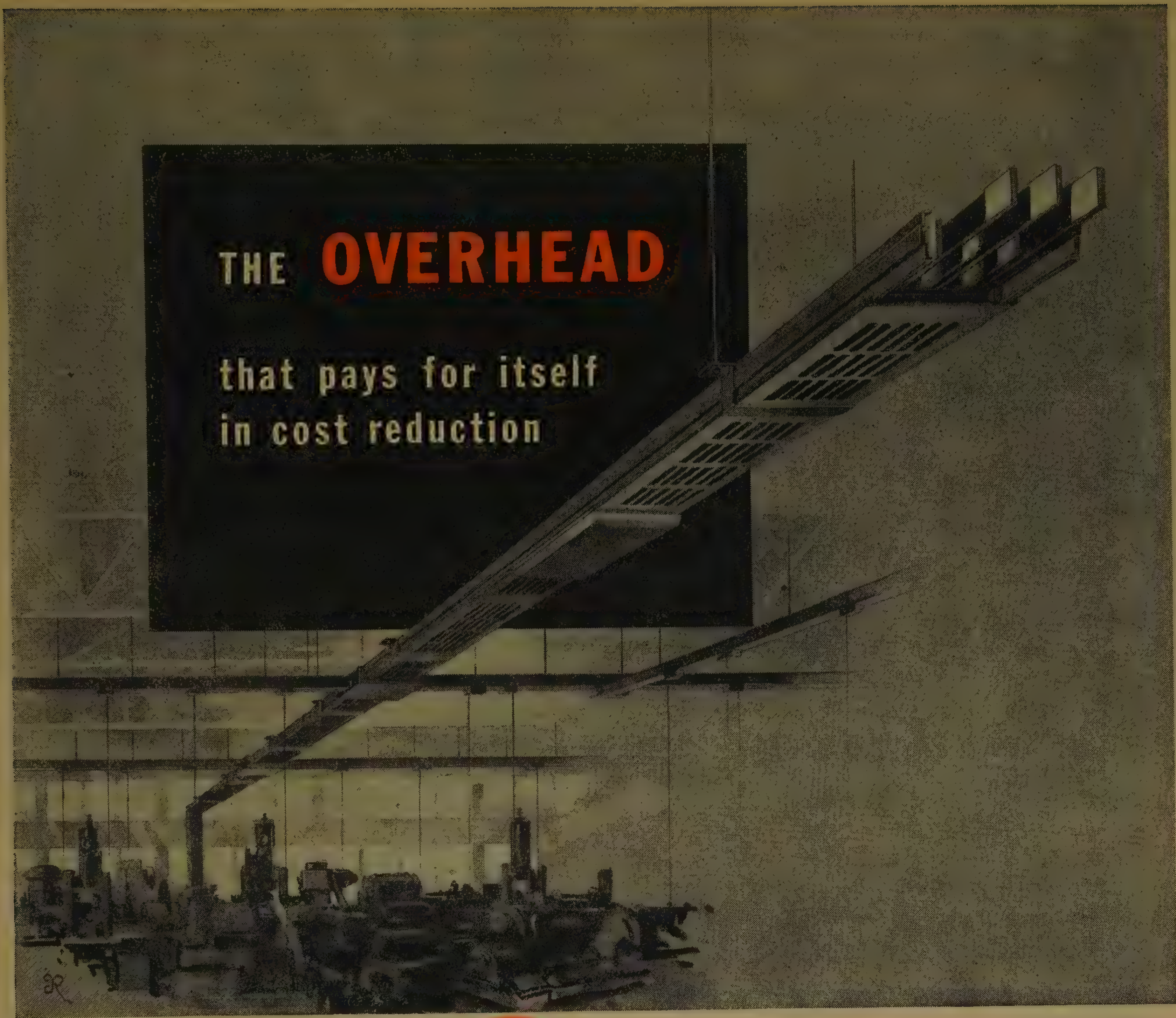
PEOPLE who follow our advertisements or pop in on us at Lansdale often say, "Why so much equipment just to test a spring?" Good question!

Making and testing a few springs is one thing. Making and testing hundreds, thousands or millions of springs is something else again. A very different something else if you want the kind of springs that can breeze by close specifications.

In order to implement our statistical control of quality (so effective in insuring better springs) . . . in order to give you springs that really meet tolerances and to *know that we know* they are right . . . we found it necessary to conceive, design and then even to build our own testing machines.

We show you a few of the machines on these two pages, some mechanical, some electronically operated.





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FEEDER DISTRIBUTION SYSTEMS

In exactly the same way that accidents, road blocks and costly maintenance plague the highway department when a road system is inadequate—so do plant managers pay for inadequate electrical power distribution.

Machines cannot efficiently operate at less than their rated voltage—overloaded lines in your plant mean costly patching of your distribution system—MONEY OUT OF YOUR POCKET.

Your inadequate distribution system is no doubt making you pay now for a full-powered, low-cost Trumbull Feeder Distribution System.

Trumbull Buss-Wa Systems (main feeders) Flex-A-Power (branch feeders) and Control Centers (motor control distribution) are available to you NOW.

Write directly to Trumbull Electric Mfg. Co., Plainville, Conn., for further information.

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WANTED—A LIGHTING ENGINEER to design and develop various types of lighting equipment. Man with some experience preferred. Will consider recent graduate of top caliber. This is opportunity for permanent connection with a firm with a record in development work for a period of over 40 years. Address, Box 411, Electrical Engineering, 33 West 39th Street, New York 18, N. Y.

WANTED—FRACTIONAL-HORSEPOWER AIRCRAFT TYPE MOTORS. Interested in buying your stocks of geared-head and direct-drive fractional-horsepower d-c aircraft type motors, particularly units manufactured by Black & Decker, Lamb Electric, Bodine, Delco, Emerson, and Leece-Neville. Indicate make, quantity, specifications, and price. Address, Box 412, Electrical Engineering, 33 West 39th Street, New York 18, N. Y.

WANTED—Prominent aircraft company, eastern United States, needs men with these qualifications. Experienced engineers with bachelor's, master's, doctor's degrees in electrical engineering, physics, mathematics. At least two years' experience in design and development of radar and television systems, automatic computers, servomechanisms, target seekers, etc., required. Positions open for preliminary and detail design, research, and development of guided missiles under Army and Navy contracts. Starting salary commensurate with experience. Address, Box 413, Electrical Engineering, 33 West 39th Street, New York 18, N. Y.

WANTED—ELECTRICAL TRANSMISSION ENGINEER. Experienced graduate engineer, capable complete design of high voltage transmission lines, including lightning protection, economic selection of voltage, conductors and structures, and preparation of estimates of cost, specifications and plans for installation. Qualifications and experience record should be submitted with letter of application. Address, Box 405, Electrical Engineering, 33 West 39th St., New York 18, N. Y.

WANTED—TWO Jr. ELECTRICAL ENGINEERS. Single men preferred. One for Electrical Transformer Engineering and one for Laboratory and Resistance Welding work to train as Resistance Welding Engineers. Address, Box 407, Electrical Engineering, 33 West 39th St., New York 18, N. Y.

WANTED—ELECTRICAL PROJECT ENGINEER. One (1) experienced graduate engineer, responsible charge electrical engineering and design of steam-electric and hydro-electric power plants and high voltage substations. Applicants please furnish personal data and experience record. Box 122, Church St. Station P. O. New York 6, N. Y.

WANTED — UNDERGROUND ENGINEER. Graduate electrical engineer, with at least eight (8) years' experience in design of underground distribution and transmission systems. Applicants should be capable of selecting most suitable types of insulations and coverings for underground cables. Experience with underground and submarine cables above 33 kv desirable. Please furnish experience record and personal data with application. Address, Box 406, Electrical Engineering, 33 West 39th St., New York 18, N. Y.

ELECTRICAL ENGINEER, B.S.E.E. 1939, M.A. Commerce 1946, age 31, 2 years field sales engineering, $1^1/_2$ years construction, 1 year research, $2^1/_2$ years Navy. Business training and experience excellent. Desire location West Coast or Colorado. Available January 1, 1947. Address, Box 424, Electrical Engineering, 33 West 39th Street, New York 18, N.Y.

ELECTRICAL ENGINEER, 15 years experience doing engineering on overhead transmission and distribution lines, $1^1/_2$ years experience as assistant supervisor of overhead line construction department. Would like responsible position with eastern utility or industrial company doing engineering on overhead lines. Registered E.E. in Ohio. Address, Box 422, Electrical Engineering, 33 West 39th Street, New York 18, N. Y.

EXECUTIVE AND ENGINEER, retired reserve officer, age 46, with vast experience in the electrical contracting business desires position of responsibility with electrical contracting or engineering firm in the Los Angeles area as manager, field engineer, superintendent, or similar position. Moving from New York to warm climate because of service connected disability. Associate member AIEE. References. Address, Box 426, Electrical Engineering, 33 West 39th Street, New York 18, N. Y.

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PROTECTION ENGINEER, broad technical and professional background, extensive experience utility system protection, planning, operation, desires permanent position in executive capacity with electrical utility company. Address, Box 397, Electrical Engineering, 33 West 39th Street, New York 18, N. Y.

TECHNICAL SERVICE SPECIALIST. Management note—unusual economical service by free lance engineer, attractive to new and progressive concerns, are offered by engineer with 18 years technical service experience in foremost concerns. Interests lie in materials, products, tests, developments, instruments, measurements, surveys, and electrical installations. Headquarters and facilities available in a co-operative but confidential laboratory in New York. Will travel. Address, Box 421, Electrical Engineering, 33 West 39th Street, New York 18, N. Y.

WANTED—GRADUATE ELECTRICAL ENGINEER by firm located in North East Ohio. Should have 2 to 3 years electrical engineering experience, preferably in appliance field. Work would consist of laboratory activities including motor development. Reply giving age, education, experience, salary desired and recent photograph. Address, Box 423, Electrical Engineering, 33 West 39th Street, New York 18, N. Y.

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COMPANY specializing in the development of electromechanical equipment has a permanent supervisory position open for engineer-experience in electromechanical development and design. Location: Minneapolis-St. Paul, Minn. Address, Box 430, Electrical Engineering, 33 West 39th Street, New York 18, N. Y.

DEVELOPMENT ENGINEER. A large manufacturer of electric and electronic devices has openings in New England for electrical engineers with 5–10 years experience in development work on vacuum tube design, cathode structures, emission characteristics, and so forth. Early interviews will be granted qualified applicants furnishing full particulars regarding age, education, experience, and salary requirements. Address, Box 431, Electrical Engineering, 33 West 39th Street, New York 18, N. Y.

ELECTRICAL ENGINEER graduate for work in chemical plant, northern West Virginia. Ninety per cent design, ten per cent field. Subsidiary of large, well-established corporation. Address, Box 432, Electrical Engineering, 33 West 39th Street, New York 18, N. Y.

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ELECTRICAL ENGINEER. Graduate electrical engineer, experienced in economic studies and design of overhead and underground distribution systems of public utilities, wanted by large consulting firm located in New York, N. Y. Applicant should submit complete information including personal data, education, and experience record. Address, Box 434, 33 West 39th Street, New York 18, N. Y.

LOUDSPEAKER for custom and domestic sound installations with performance comparable to large theater types now developed through pilot model and limited production stages including patent protection, needs plant, market development, and organization beyond my means. Would consider forming, joining, or being joined by existing organization or individuals to form team to exploit this and other ideas. Communicate directly with Paul W. Klipsch, Hope, Arkansas.

PHYSICISTS AND ELECTRICAL ENGINEERS. Unusual opportunities in recently organized research group for work in broad fields of radio, radar, telemetering, servomechanisms, and gyroscopes. Excellent openings for experienced physicists with good training in fundamentals. Please reply, giving experience and education including transcript of college record, to: Engineering Personnel Office, NORTH AMERICAN AVIATION, INC., Municipal Airport, Los Angeles 45, Calif.

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GRAD E.E.; age 40; 7 yrs util, 5 yrs constr and 3 yrs Army exper. Dependable, loyal, sober; desires sales pos. Now available. Will travel. E-257.

ELEC ENGR; 12 yrs exper—elec util distr system des, indus elec des, operating reports and studies, appraisals, pub util management. Excellent background electronics. Organizational exper. Prefers milder climates. E-258.

RECENT GRAD in elec engg (power) from Columbia Univ desires employment. Training in pwr transmission, distr, and utilization—familiar with symmetrical components. Served as Ensign U. S. Navy. E-259.

ENGR with record of commercially successful creative thinking and invention. Doctor's degree from Cornell in chemistry and physics; 20 yrs exper in elec, mech, chem engg. Research, devpmt and des. E-260.

JR ELEC ENGR; B.S. from Tufts College, 1945; desires pos with electronic or machinery mfg firm. Served in U. S. Navy for three yrs; 8 months with rank of Ensign. E-261.

ELEC ENGR; B.S.E.E.; 27, single; 5 yrs exper in test and devpmt. Desires permanent pos with oppor for advancement. E-262.

ENGR; B.S. degree in Comm and Pwr; 6 months experience radar, 18 months point-to-point comm; prefers pos in television or other high freq work. E-263.

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ELEC ENGR; B.S.E.E.; 25, single; 4 yrs exper in supervising elec des, constr, operating pwr plants, substations, transmission and distr systems. Has administrative background. Fluent Spanish. Consider anything with future. Will go anywhere. Will travel. E-265.

Positions Available

ASSISTANT PROFESSOR of electrical engineering, to teach power generation and distribution. Salary, about $3,600 per academic year. Location, North Dakota. W-7606.

ELECTRONIC DESIGN ENGINEER, either electrical engineering training or a physicist who has specialized in electronics. Should have had experience in research and development work. Salary, $4,000–$6,000 a year. Location, Rhode Island. W-7634.

ELECTRICAL OR MECHANICAL ENGINEER, 35–45, with design and production experience on small transformers, to supervise project for fluorescent manufacturer. Salary, $5,000–$6,000 a year. Location, Ohio. Interviews, New York, N. Y. W-7641.

PROFESSORS of electrical engineering, Ph. D. or D. Sc. degree, 35–45, with considerable experience instructing radio engineering, including broad experience in use of radio equipment and apparatus; some experience with commercial organization as design or development engineer on radio equipment and with background and training in electromagnetic theory. Location, upstate New York. W-7678.

INSTRUCTOR, electrical graduate, to teach direct and alternating current theory and machinery laboratory, and probably electricity and magnetism. Salary, $3,000 a year. Opportunity to teach evenings. Location, Pennsylvania. W-7814.

TECHNICAL WRITER, 26–35, electrical graduate or equivalent, with three to five years' experience, preferably in instruction, maintenance and sales manuals, etc., to prepare for final approval. finished copy and layouts covering technical information relating to mechanical, electrical, electronic and hydraulic devices and equipment. Salary, approximately $4,480 a year. Location, northern New Jersey. W-7877.

ELECTRICAL ENGINEERS, graduates, with an electronics major, for electro-mechanical developing laboratory, to design and develop electronic circuits for airborne fire control equipment involving servomechanisms, amplifiers and industrial application of electron tubes. Must have at least four years' experience in development of fire control circuits, radio receiver design or closely allied field. Knowledge of stabilization and sighting systems with radar control of ranging or gun laying preferred. Salary, approximately $4,200 a year. Location, New York, N. Y. W-7934.

PROJECT ENGINEER, electrical graduate, to work on sound recording equipment. Should have ten years' experience in development and design of audio amplifiers for sound or speech recording equipment. Knowledge of production methods on this equipment desirable. Salary, $4,680 a year. Location, New York, N. Y. W-7996.

APPLICATION ENGINEERS, under 35, with some experience in vibration control, rubber or plastic molding. One to do general sales, one for general all-round mechanical, one for electrical application, and one for special motor application. Salary open. Location, East. W-8015.

ELECTRONIC INSTRUMENT ENGINEER, B. S. degree in a recognized school in electrical engineering, preferably with electives in the fields of communications or instruments. Must have had two to five years' experience in the development or use of electronic instruments. General experience in the field of electronic circuit development or application is also desirable. Salary, $3,600 a year. Location, Connecticut. W-8045.

ENGINEERS. (a) Engineering Assistants, 26–40, electrical graduates, with minimum of four years' engineering experience, preferably in electrical utility. Work is in conjunction with field and office layout and design of transmission and distribution system, substations and power plants, system planning and protection, estimating, etc. Salary open. (b) Cadet Engineer, 21–25, electrical graduate, for two years training in various operating departments of electrical utility leading to engineering position in layout and design work on transmission and distribution systems. Starting salary, $175 a month, increase after six months. Location, Delaware. W-8057.

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CONSTRUCTION AND OPERATING FEATURES

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2. High interrupting disconnect fuses, hook stick operated.
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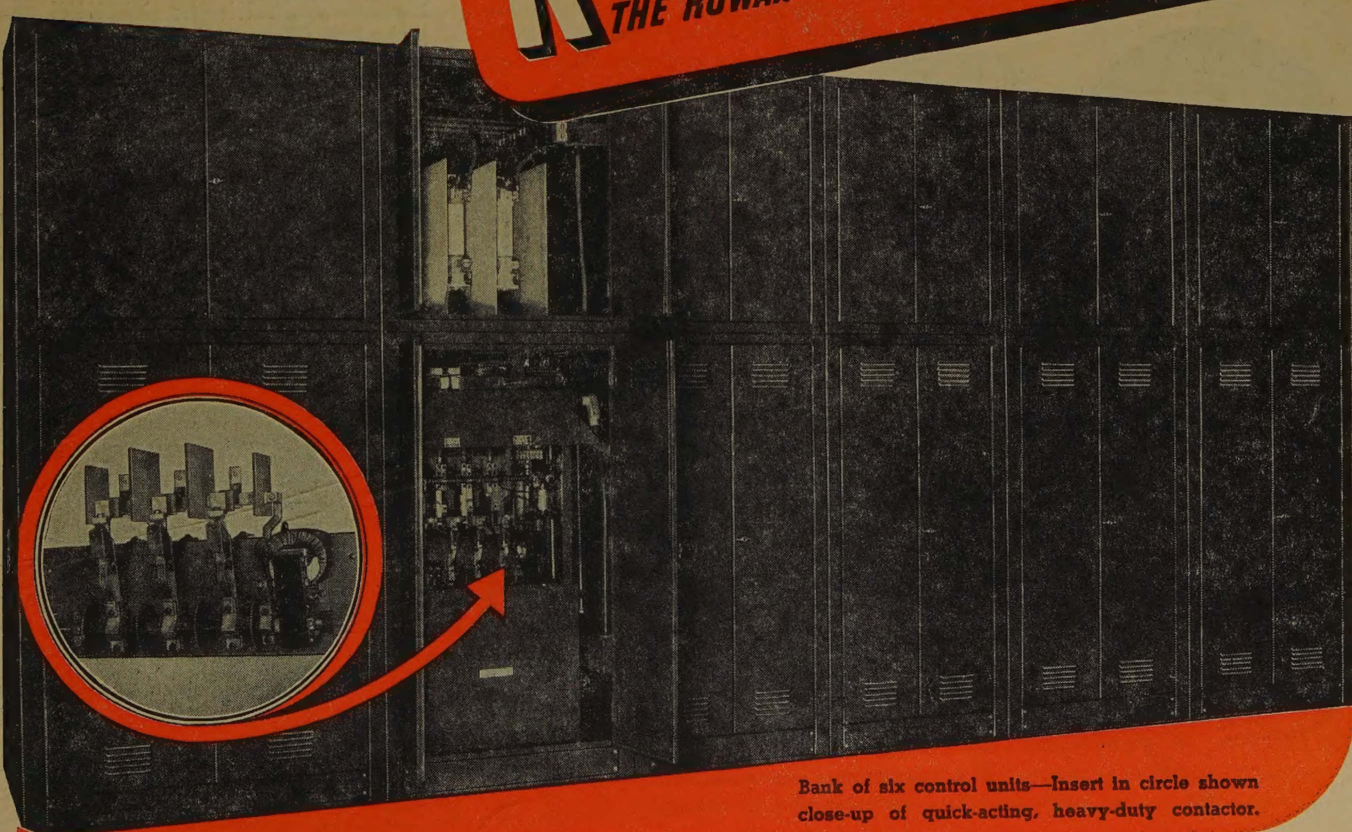
ROWAN CONTROL

THE ROWAN CONTROLLER CO., BALTIMORE, MD.



Bank of six control units—Insert in circle shown close-up of quick-acting, heavy-duty contactor.

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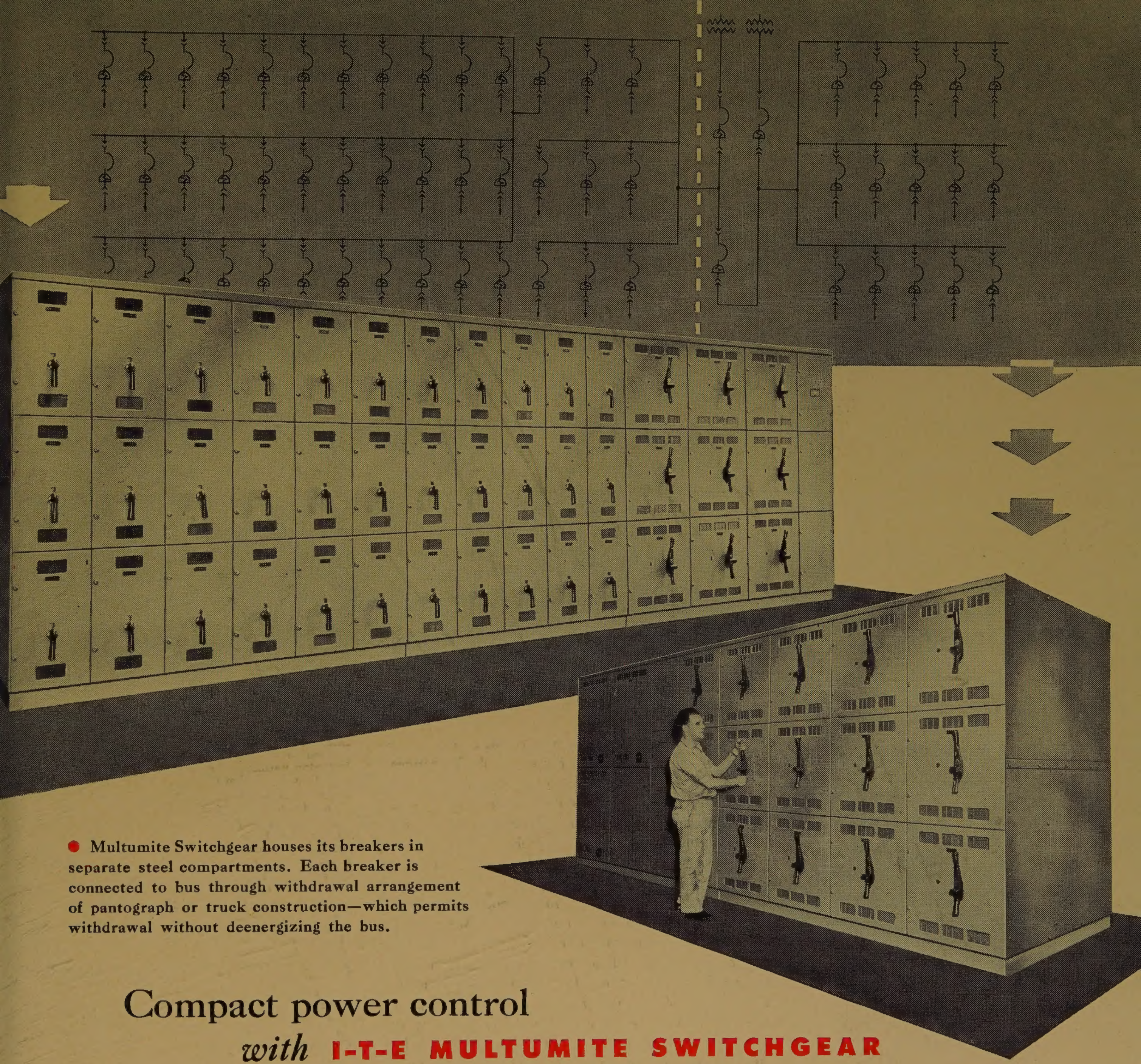


● Multumite Switchgear houses its breakers in separate steel compartments. Each breaker is connected to bus through withdrawal arrangement of pantograph or truck construction—which permits withdrawal without deenergizing the bus.

Compact power control
with I-T-E MULTUMITE SWITCHGEAR

These units control 460V plant operations at a new non-ferrous rod rolling mill in the mid-west.

And, they're completely factory-assembled.

Diagram shows what's inside of the two sections which, actually, are joined as one.

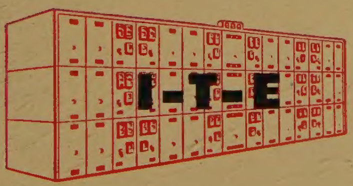
Power, from 1500 KVA transformers, leads into the bus-tie and sub-main circuit breakers in the unit on the right. Then it is distributed to plant operations through the feeder circuit breakers in the unit on the left.

Continuity in every one of the mill's precision operations—pumps, compressors, annealing furnaces, drawing dies, etc.—is insured with one of the elements in the centralized power package.

For complete technical information on "Multumite Switchgear Assemblies", send for Bulletin 4208, I-T-E Circuit Breaker Co., 19th & Hamilton Streets, Philadelphia 30, Pa.

I-T-E

HIGH SPEED CIRCUIT BREAKERS • SWITCHGEAR • BUS STRUCTURES • SUBSTATION UNITS • AUTOMATIC RECLOSING CIRCUIT BREAKERS



MULTUMITE SWITCHGEAR

SOLITE* CAPACITORS

new! startlingly smaller!

ACTUAL SIZE 1 MF, 200 WVDC

TODAY'S trend to ultra-compact electronic designs strides forward with this remarkable, new Solar development in truly tiny paper capacitors.

*SOLITE** Capacitors, utilizing a unique, revolutionary self-healing metallized paper construction, occupy approximately one-third to one-fourth the volume of equivalent conventional multi-paper capacitors. There is a similar saving in weight.

A comparison of typical ratings in tubular types tells its own story:

Dimensions in Inches							
Capacitance	WVDC	SOLITE		Conv. Design		Weight in Ounces	
(mf)		Length	Diam.	Length	Diam.	SOLITE	Conv. Design†
0.1	200	5/8	3/8	1 5/8	1/2	0.08	0.32
0.5	200	1 1/8	15/32	2	11/16	0.13	0.75
1.0	200	1 1/8	17/32	2 1/2	13/16	0.26	1.15
1.0	400	2 1/8	11/16	2 1/2	1	0.69	1.75

† Based on aluminum foil construction. Lead foil capacitors will be still heavier.

*SOLITE** Capacitors are available in both non-metallic and metallic housings in standard d-c voltage ratings up to 400 volts. *SOLITE** Capacitors are also supplied for alternating current applications.

Pilot quantities of *SOLITE** Capacitors may be had immediately. Solar is prepared to discuss delivery schedules of production quantities for your use in those specific applications where you can take best advantage of this important new advance in the capacitor art.

Full details of *SOLITE** Capacitors may be obtained on letterhead request from: Solar Manufacturing Corporation, 285 Madison Avenue, New York 17, N. Y.

"WHEN SPACE IS TIGHT, USE *SOLITE*"

* Trade Mark

Solite Capacitors are fully protected by U. S. letters patent and patents pending.

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SOLAR CAPACITORS

"Quality Above All"